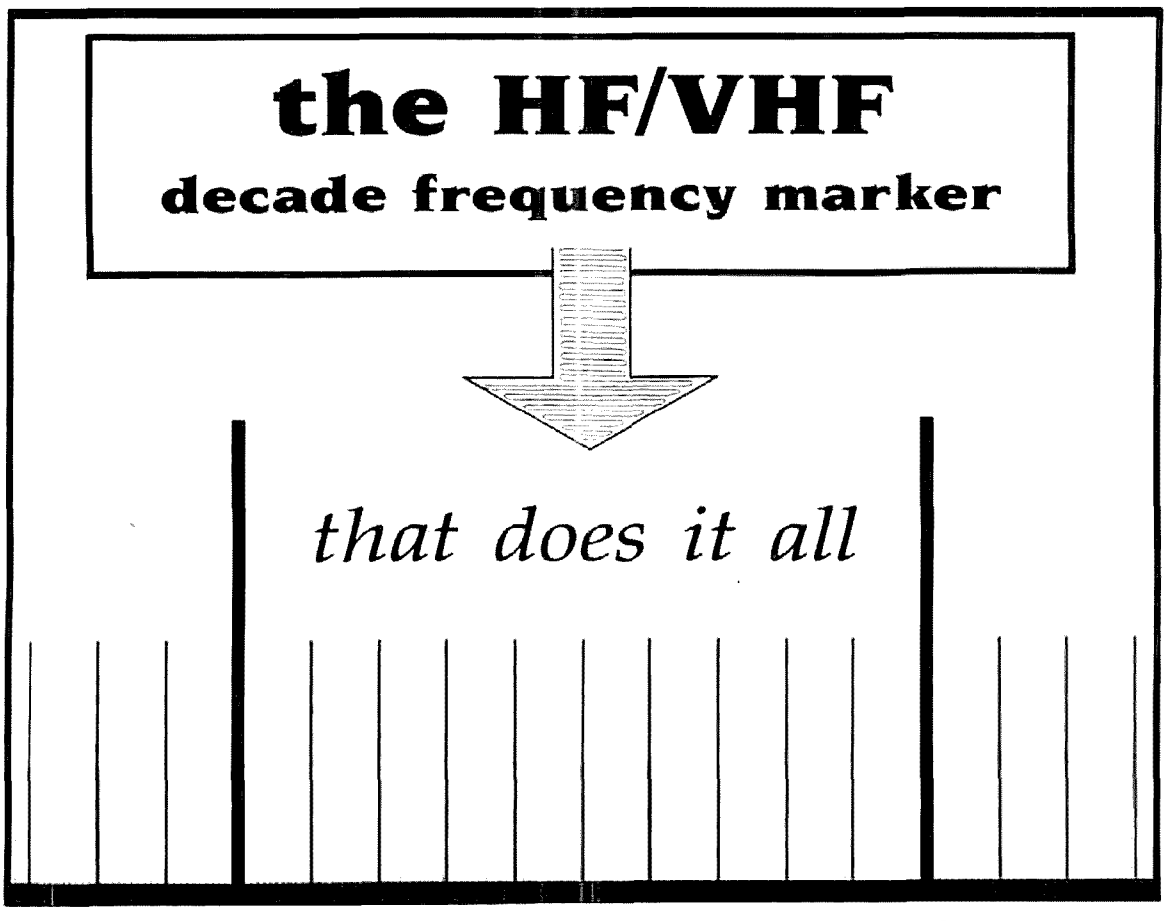
ham radio magazine

the HF/VHF
decade frequency marker



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on
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publisher

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editor-in-chief
and associate publisher

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assistant editor

Joseph J. Schroeder, W9JUV
Alfred Wilson, W6NIF
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Rally Dennis, KA1JWF
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Susan Shorrock
circulation manager

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circulation

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REFLECTIONS

before you panic, get the facts

Picture yourself in a quiet restaurant with your spouse or friend. Without warning, the peaceful atmosphere is disturbed by a loud discussion — its content quite private in nature — from across the room. It gets your attention and you find yourself listening. But suddenly you're jolted by an authoritative voice directed at *you*, informing you that you've just broken the law. Your offense: *listening to a private conversation*.

That scenario, half-jokingly told to me by Joe Schroeder, W9JUV (see *Presstop*), has a remnant of truth in an analogous situation. There are presently in the works two bills (HR-3378 in the House of Representatives and S-1667 in the Senate) that could make it a federal crime to listen in on certain conversations received over the airwaves.

Before you panic, listen to the following facts. Several voice and data-handling organizations — specifically the Cellular Telecommunications Industry Association (CTIA), the Electronic Mail Association, Telocator, paging, and video conferencing industries — do not want their communications intercepted. Sounds reasonable enough. The problem, as I see it, is in their methods. These groups are lobbying (quite strongly, I might add) for Congress to pass what is now known as the Electronic Communications Privacy Act of 1985, which would prohibit anyone in the United States who has a general coverage receiver or similar type device from listening to transmissions at will. If, for example, you inadvertently tuned your receiver or scanner across a cellular phone conversation and *continued* to listen, you'd be breaking the law.

The problem is not the desire of these industries to protect their customers' privacy, but their assumption that the general public should be asked to forfeit its right to listen to radio transmissions without restriction.

What about us Radio Amateurs? Aren't we excluded from this situation? Yes and no. In preliminary hearings, six broad categories — including Amateur Radio — were considered. But when the subject of Amateur autopatches came up, an initial response from one of the subcommittee members was that since the telephone is involved, reception of these types of transmissions might not be part of the exclusion.

Keep in mind that both *the* House and *Senate* subcommittees drafting this act — which, incidentally, may turn out to be the single most important piece of legislation affecting us since the 1934 Geneva Radio Convention accords — are still in the preliminary phases of the project. *This* is the time to let your senators and representatives know that you're very interested in these bills even before they come out of the Senate and House subcommittees.* You want your representatives to acquire the facts and understand your concern about potential encroachment on and abridgement of your operating privileges. It might not hurt for you to remind them of the helpful services Radio Amateurs have provided and will continue to provide to the public in emergencies — without desire for or expectation of recompense. It would be nice to leave them with some positive suggestions that they could take to the subcommittees drafting the legislation.

The problem, as mentioned before, is one of fear: people in the commercial communications industries are concerned about possible interception of their signals. But as foolish and unworkable as it would be for the "authorities" to tell all the clients in the hypothetical restaurant scene to close their ears, so would it be for the government to tell radio owners what kinds of transmissions they should or should not listen to.

What would make more sense would be for the cellular and other voice and data-handling organizations to utilize some form of encryption — be it non-standard forms of modulation, tone insertion, inversion, or coding. The necessary technology exists, and one application of it is in daily use in police departments across the country. Otherwise, just imagine a scene in a penitentiary, wherein one pathetic soul is asked by his cellmate, a thrice-convicted murderer, what *his* crime was. . . and the poor soul replies, with great timidity, that he was caught listening to a radio transmission.

Rich Rosen, K2RR
Editor-in-Chief

*In the House of Representatives, the Subcommittee on Courts, Civil Liberties, and the Administration of Justice; in the Senate, the Subcommittee on Patents, Copyrights, and Trademarks.

presstop de W9JUV

A SERIOUS POTENTIAL THREAT TO ESTABLISHED AMATEUR RADIO PRACTICES may be developing as the "Electronic Communications Privacy Act of 1985" moves toward consideration by both the U.S. Senate and House of Representatives. The two bills, HR-3378 in the House and S-1667 in the Senate, were initially intended to simply make earlier Congressional action against wiretapping apply to radio-relayed data transfer. They provided an opportunity that was quickly seized by the cellular telephone industry, already deeply concerned about the security of its 800 MHz communications as scanners move into the microwave region.

Although Amateur Radio Is One Service Specifically Exempted, along with fire, police, and several others, from the bills' prohibitions against unauthorized "interception," its unrelenting "protectionist" character could drastically change, if not prohibit, some popular Amateur services. For example, the bills' drafters would apparently characterize any "interception" of a radio-relayed telephone conversation by anyone except those actually participating as a federal crime -- with possibly no exception for Amateur autopatches or phone patching! One subcommittee staff member has even been quoted as feeling that all Amateur interconnection with the telephone system is possibly illegal.

Though the Premise -- Privacy For Radio-Carried Personal Communications -- is a worthy one, the bills could, in effect, result in a case of classic overkill. In its present form, the proposed legislation could, for example, declare vast portions of the RF spectrum, including most of the short and long wave bands, illegal territory for the SWL or other radio hobbyist -- not to mention the tens of millions of scanner owners in the U.S. Thus far the League's position has been one of "wait and see," in apparent belief that Amateur Radio's "exemption" will be sufficient to stave off any negative effects. However, in the absence of strong Amateur representation in protection of current Amateur privileges and practices, the bills' impact on Amateur Radio could actually be quite severe.

Amateurs Should Contact Their Senators and Representatives and advise them of their deep concern for the potential effects of these bills on both Amateur Radio and other citizens' activities. Offers of technical expertise as to the legislation's effects might be welcomed by the legislators as well. Be sure to insist that they keep you advised as to progress and significant changes as the bills move toward passage.

RADIOLOCATION HAS BEEN ASSIGNED THE 1900-2000 KHZ BAND, the FCC announced in early November. Amateur Radio will, however, continue to share that portion of 160 meters on a Secondary basis. Impetus for the shift was WARC '79, which expanded the AM broadcast band upward into Radiolocation's former 1605-1705 kHz slot. Though a "Secondary" status means Amateur operators must take pains to avoid interfering with operations of the Primary users, the net effect on Amateurs should be negligible.

Users Of Wideband Radiolocation Systems Could Be Licensed on 160 as early as December, 1985, but no large-scale movement to their new band is expected until mid-1987 -- and even then usage is still likely to be light and mostly in limited geographical areas.

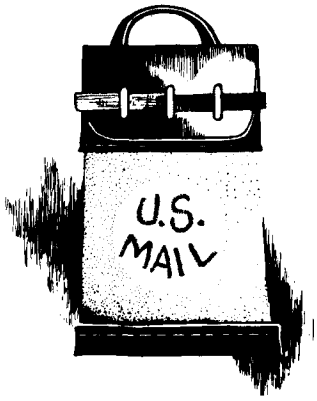
A CONFERENCE TO DISCUSS PROGRESS AND PROBLEMS OF THE VOLUNTEER EXAMINER PROGRAM is being planned during the Orlando Hamfest, March 7-9, 1986. Sponsored by CARE (Council for Amateur Radio Examining), the session will be open to all VECs, VEs, and other interested parties whether CARE members or not. Contact CARE for further details.

CARE Membership Rates Have Now Been Set, and all who've requested information on the new national VEC organization should have received membership information by now. Dues for a VEC organization are $25 per year, while General members (VEs accredited by a member VEC) must pay $10 annually. Dues for At-Large members (VEs with accreditation from a non-CARE member) and Associates (anyone else) are $15.

CARE Now Has A Permanent Mailing Address: P.O. Box 866, Glenview, Illinois 60025.

AMATEURS WHO'VE LOST THEIR LICENSES MAY NOT USE ANOTHER AMATEUR'S STATION for any further Amateur operation, the Commmissioners decided in a Report and Order on PR Docket 85-51. The question had arisen in the case of a former West Coast Amateur whose license had been revoked for various violations and who wanted to remain active using friends' stations. The ban pertains to any Amateur licensee whose license has been revoked or suspended for all or part of the license term, who has surrendered his license for cancellation in connection with an action against him, or who is subject to a "Cease and Desist" order that relates to his Amateur Radio activities.

EXTENSIVE RESTRUCTURING OF CANADA'S AMATEUR LICENSING has been proposed by Canada's Department of Communications. The new no-code entry-level license would provide full privileges above 30 MHz (250 watts maximum) using only commercial transmitting equipment; the exam would cover only station installation and operation, very basic theory, antennas, and rules. Moving to the second level would require only a 12-WPM code exam and would extend the same privileges below 30 MHz. Top level would require passing an exam on "advanced electronic theory," then permit using home-built equipment and up to a kilowatt.



comments

Novice Privileges

Dear HR:

With reference to T. Dillingham's (KAØDOE) comments concerning Novice privileges, ("Comments," September, 1985, page 15), it seems KAØDOE wants a "reward" for being a Novice for six years. This is itself a reward, as originally the license was a one-year, non-renewable affair, and before that, of course, did not exist at all.

I can't believe that anyone who aspires to increased operating privileges can't master the skills necessary to pursue their interests in a six-year period.

The tone of this letter suggests that, for some at least, adding any additional benefits would only make them more complacent, thus defeating the purpose.

Ray Cunningham, W3YBF
Ambler, Pennsylvania

Dear HR:

We are in sorry shape, I fear. A casual chat with our Section Manager revealed that many Amateurs in Nebraska are against the ARRL's enhanced Novice Privileges Proposal — some adamantly so. If this is true throughout the United States, I think we have big problems in that I believe this is just the attitude that will be the death of our hobby. Look at the numbers on page 9 of the August, 1985, issue of *QST*. The numbers are *still* not in our favor.

One of the favorite arguments seems to be that the bands are crowded enough already. Oh, really? Has anyone tried operating the HF bands on a weekday mid-morning? People who work nights should be made a target of our efforts if there is a general feeling that the bands are too crowded. The bands certainly are not overpopulated most weekdays.

Some repeater operators are concerned about losing 220 MHz to Novices because they may have to find new territory for their control links. Yes, there are lots of repeaters...probably more than we need in many areas, although not in Nebraska. But if they could take away band privileges by state they could easily revoke 220 here since the control links wouldn't even occupy a space of several kHz, let alone 5 MHz. We are not going to keep the band if we just use it for control links. We must get more people on the band one way or another. A space can be reserved for the control links by band plan as is done now. Novices can respect band plans, too: they must on the HF bands, and do.

As one who acquired Novice privileges when phone operation was permitted on 2 meters, I see no problem in once again permitting Novices to operate phone. This, as has been stated in League literature on the subject, is even less of a problem now since the Novice license is renewable. During my term as a Novice, the phone privileges were taken away. I have often wondered whether this was a contributing factor to the seemingly low numbers of hams my own age (32) and other hams licensed about that time (1966) on-the-air.

It's time to wake up, folks. There will still be plenty of room for those of us who like to tinker and prefer CW and SSB ragchews. But we have a dual problem: lack of population on the ham frequencies will lead to their loss, and ultimately to the loss of the hobby; the lack of population will become worse if would-be newcomers are not interested in joining our ranks. This is the only real stab at a solution yet proposed and it appears to be one which has been given much forethought. Hams spoke out against no-code. Here is a compromise. Let's put away petty differences and get behind it for the continuance of our hobby.

Michael S. Lennen, KDØEV
Omaha, Nebraska

to key or not to key?

Dear HR:

Print this if you like. It's just for fun.

To Key or Not To Key

To key or not to key
— that is the question.
Should we drop the code exam?
Dismiss the simple switch
Or retain the day when ability
was reckoned by the fist?

To master the key
is to be
a radioman.
To scorn the key
is not
to be.

2001 — a CW odyssey.
What a sci-fi mode!
For each Jedi knight,
a practice oscillator
was dutifully stowed.
Even on the Enterprise
Uhura copied code.

Does the future hold
digital radio?
Eighty-meter beams on
a chip?
Or will we turn 'round
to see
Digital radio
to be
the key?

Bob Zavrel, W7SX
Sunnyvale, California

a frequency and level standard

A little black box and some digital pulse magic produce a frequency-stable, high-level generator

By Hans Evers, PAØCX, Am Stockberg 15, D-5165, Hürtgenwald, West Germany

A good RF signal generator needs an effective electrical shield around its generating circuit. Without such a shield, it would not be possible to guarantee an accurately determined output level over a wide range of frequencies. In other words, the output attenuator would lose its credibility.*

Shielding an RF circuit requires more than just placing the components behind a metal wall. Leakage and radiation will occur from almost anything electrical that happens to protrude from the circuit enclosure. Sometimes a rather sophisticated filter system is required for stopping the interference that would otherwise spoil the output-level calibration, particularly at low levels and at high frequencies.

This shielding requirement is only one reason a good RF signal generator is usually an expensive piece of test equipment, and difficult to build yourself. For me, it was *the* reason behind the rather unusual design of the frequency and level standard, which, of course, is just another name for a signal generator.

design concept

In principle, the idea is simple. A circuit contained in a hermetically sealed box can't have any filter problems if there's nothing to filter in the first place. That's why I thought of designing a signal source without external controls, power cable, tuning, or level meter — just a box with an output connector. Everything except the output is kept inside.

Thinking a bit further, putting such a drastic principle into practice was not as easy as it sounds, and a few concessions were unavoidable here and there. For example, it's difficult to tune an oscillator and to have its frequency displayed through a solid shield. Consequently, I had to restrict the output signal to cover only a certain number of predetermined frequencies.

The simplest way to obtain many fixed frequencies simultaneously is by adopting the frequency-marker principle; that is, by deriving harmonics from a crystal oscillator. This also guarantees good frequency stability. To keep the many signals separate from each other they must be labelled; I gave them, therefore, different amplitudes, making them clearly distinguishable. This was done with the help of a very elementary digital processing circuit.

A simple action — such as flipping a switch — could pose a problem if it had to be done through a metal shield. But this is also no problem: mercury switches mounted inside the circuit can be actuated simply by turning the box to a different position.

The part that needs most switching, the variable output attenuator, is kept outside the shielded enclosure. This allows the signal generator to produce a uniform, but accurate, signal level over a wide frequency range. In fact, it turns out that it's not only the generator itself, but also the variable attenuator that determines the eventual level accuracy at higher frequencies. And the outboard attenuator can be easily replaced by an improved version at a later date.

The cost of the active components (six ICs, including the voltage regulator) should come to not much more than $3.00. To sum it up, at least for many applications, the box-with-plug idea works the way I hoped it would: as a good signal generator that is neither expensive nor difficult to build.

synthesizer

Many people who've tried to calibrate a receiver frequency scale using a frequency marker have had the experience of being confused by the different, but

*That is to say, RF from a poorly shielded generator could combine with a highly attenuated signal and provide erroneous results — Ed.

look-alike, signals. It is therefore an advantage to introduce signals that belong to a different order in the spectrum hierarchy. First the 1-MHz markers, for instance, then the 100-kHz markers, and so on. Usually this is done by turning a switch.

To avoid using such a switch, I tried a different method. The coarse markers were labelled by emphasizing their amplitudes (**fig. 1**). After frequency division of a 1-MHz crystal-oscillator signal, pulses of different lengths arrive at the NAND gates. The product is used to interrupt the stream of 1-MHz oscillator-output pulses at regular intervals by longer or shorter spaces (**fig. 2**).

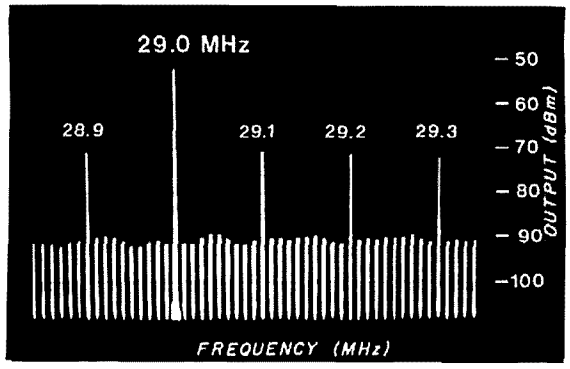


fig. 1. Typical output display spectrum as seen on a frequency-spectrum analyzer. Multiples of 100 kHz harmonics have signal strengths corresponding to the standard S9. The MHz markers are about 20 dB stronger; the 10-kHz markers are 20 dB weaker.

The underlying mechanism that produces the different phenomena may be easier to understand when you realize what the pulse sequence would look like if the polarity had been reversed. Tuned circuits in a receiver recognize certain patterns in the stream passing by and therefore respond to certain rhythms. Thus, for example, the one-pulse space in the otherwise steady flow of 1-μs (1-MHz) pulses represents, in fact, a pulse of relatively longer duration. If such an event occurs regularly every 10 μs, it is capable of causing a response in a circuit tuned to a multiple of 100 kHz (being the repetition rate of the space).

The reason amplitudes of the 1-MHz markers stand out from the rest is not difficult to see. The train of 1-μs pulses with each 10th pulse removed produces more energy in a circuit tuned to a 1-MHz multiple than to one tuned to a 0.1-MHz multiple (where, at the same time, the circuit has to manage with only one out of ten pulses).

The two-pulse spaces responsible for the 10-kHz markers (occurring every 100 μs) produce energy that must be shared over 100 times as many signals in the frequency spectrum. This explains why the 10-kHz markers are so much weaker than the others. Of

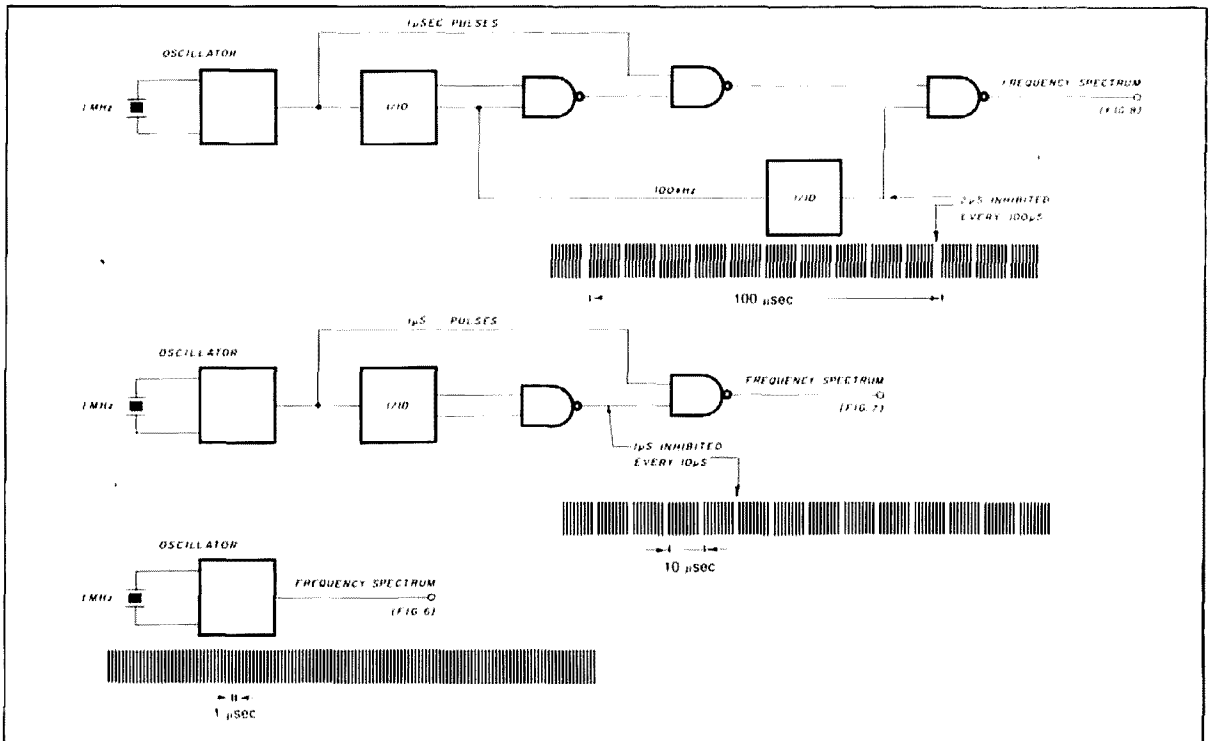


fig. 2. Method by which the frequency spectrum is synthesized by creating patterns in a digital pulse sequence.

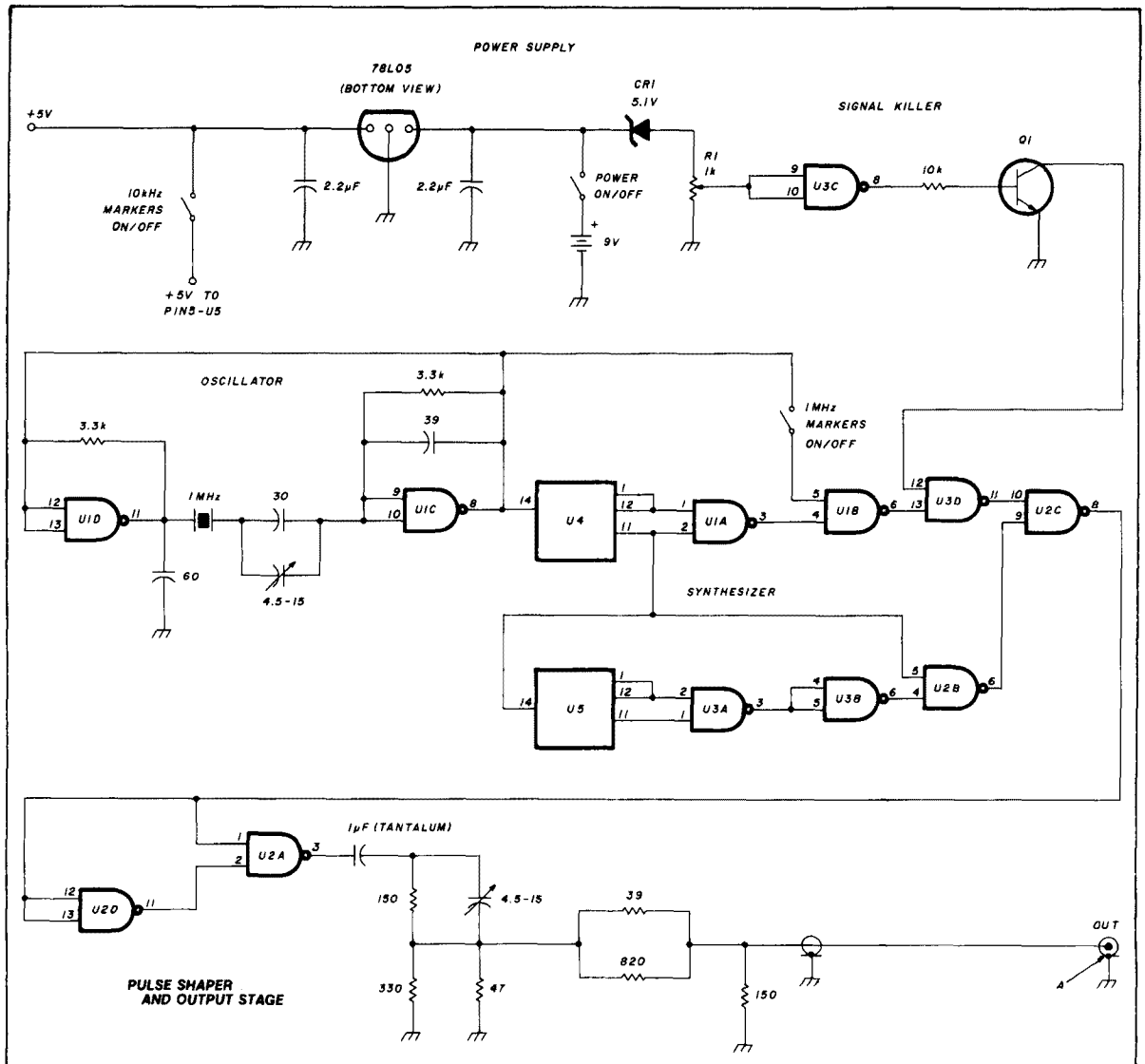


fig. 3. Schematic diagram. The entire circuit, battery included, is contained in a closed metal box which is electrically connected to the circuit ground only through the coaxial cable to the output connector at point A. U1 and U3 7400; U2 74S00; U4 and U5 74SL90.

course, all this tampering with pulses makes the total architecture rather complex. It is, therefore, not surprising to find that there's a practical limit to where the results remain acceptable, and where further refinements would create a less orderly spectrum. What happens is that if the mixture of pulses and spaces in between gets too complex, less-wanted patterns become recognizable. This results in poorly determined amplitudes. Therefore only the *100 kHz* position (only multiples, all with identical amplitudes) should be used as a level standard. The spectrum-analyzer pictures show how the amplitudes tend to become irregular as the frequency spectrum becomes more complex.

Thus the imperfection is not so much the result of what a typical analog thinker (as many Radio Amateurs appear to be) might suspect: the wiring. Digital circuits are remarkably tolerant of construction-related problems such as breakthrough or crosstalk — and, for that matter, of poor layout, messy wiring, and horrible-looking breadboard creations. What is also appreciable is that from the synthesizer output onward, only the *absolute* amplitude may change. The amplitude ratios between the different orders of frequency, once they have been encoded into a digital pattern, are not likely to be changed by following wave-shapers, amplifiers, or other analog circuits.

Figure 3 shows the circuit as it was eventually

adopted; fig. 4 shows the socket connections of the ICs. Note that an electrolytic capacitor is provided at each individual IC for additional voltage protection.

Figure 5 shows an oscilloscope display of a detail from the synthesizer output pattern showing some 1-pulse spaces (indicated by A) responsible for the 100-kHz signals, as well as the 2-pulse spacing (B) that produces the 10-kHz markers. **Figures 6, 7,** and **8** show the different frequency spectra that are created by the pulse patterns given in **fig. 2.**

output stage

The output stage consists mainly of a pulse shaper, or "harmonic producer," based on the fact that the shorter the pulse duration, the more frequencies are produced, and consequently the frequency spectrum will be wider and the harmonic amplitudes more uniform, though lower.

How short should the pulse duration be for obtaining a reasonable compromise between frequency range and output level? For lack of more specific information in the otherwise plentiful literature on frequency markers, I investigated a few circuits I found in Amateur Radio publications.

The output of a 7400-type NAND gate turns out to have an impedance of about a few hundred ohms and could, if necessary, be safely connected to a 50-ohm load. (The IC would even survive a full DC short from its output terminal to ground, if it ever came to that.) However, to permit a proper match into a 50-ohm device, such as an attenuator, a small network is inserted. This also permits an increase in frequency response at the high-frequency end of the spectrum.

Figure 9 illustrates some of the networks that I tried. Note the important improvement by the one shot in fig. **9B.** This simple circuit turns out very short pulses indeed. In this circuit, the last NAND gate receives the pulse directly at its top input. The pulse is inverted at

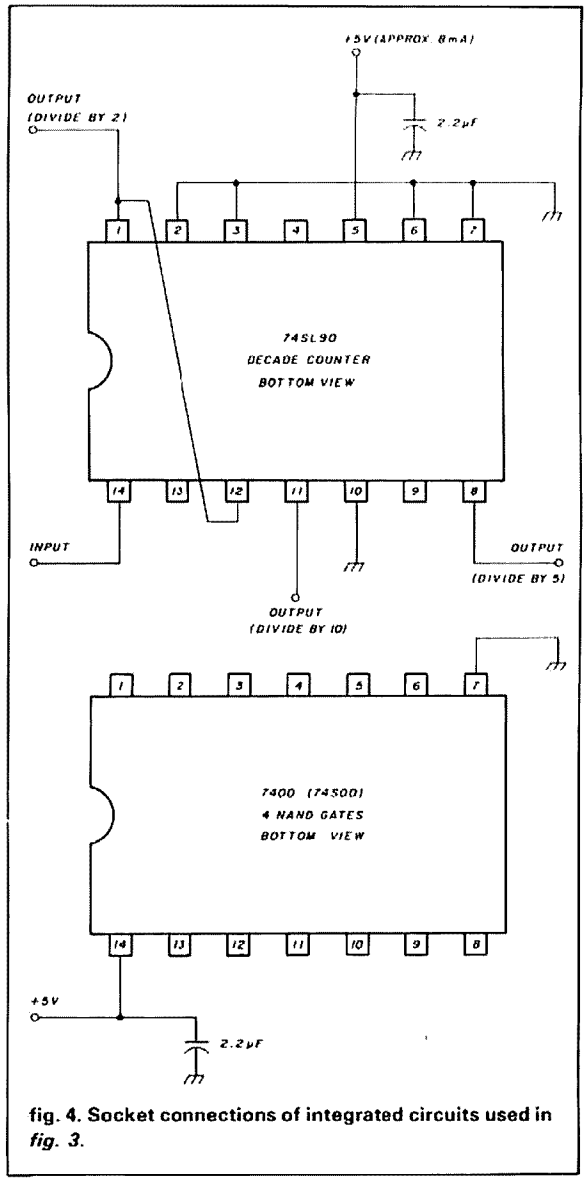
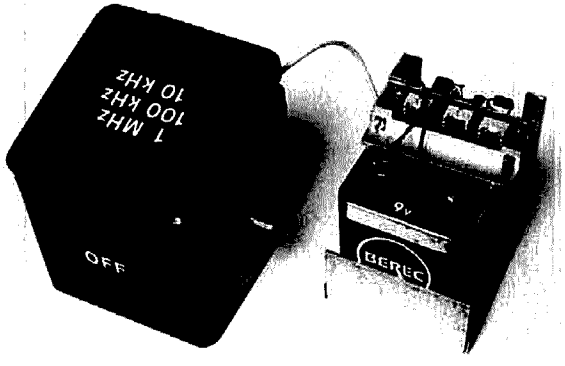


fig. 4. Socket connections of integrated circuits used in *fig. 3.*



The frequency and level standard outside its enclosure. Normally the entire circuit, power supply, and all control elements are inside the black metal box; thus all RF leakage is eliminated without the need for any filtering.

the bottom input, and is delayed only by another NAND which acts as the inverter. As the output of a NAND gate is low only if both inputs are high, the total effect could perhaps be compared with letting a signal rush through the circuit, and immediately slam the gate shut behind it, chopping its tail. Thus, using an ordinary 7400, all output pulses have a duration that can't be longer than about 10 nanoseconds, the delay time of the inverter. The circuit produces adequately short pulses for creating a reasonably flat frequency response that covers all HF bands, requiring only a slight correction.

The 7400 also exists in a more than three-times-as-fast Schottky version, the 74S00. This device produces even shorter pulses and makes it possible, at least with

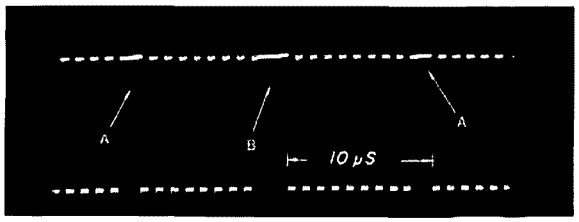


fig. 5. Synthesizer output signal as displayed on oscilloscope. At *A* one pulse is inhibited, repetition rate 10 μs. At *B* two successive pulses are inhibited, repetition rate 100 μs.

some correction again, to create a flat response that comfortably includes the 2-meter band. At the time the spectrum (which starts to roll off at 160 MHz) reaches 400 MHz, the amplitudes have dropped by about 15 dB. This would still make a good frequency marker for the 70-cm band.

In the final circuit (**fig. 3**) an extra 50-ohm, 6-dB attenuator is inserted. This puts the output level at −73 dBm and provides isolation between the frequency-correcting network and a possible capacitive load at the output connector.

The −73 dBm level was chosen for a good reason. It allows sufficient compensation of the frequency response for including the 2-meter band. It also corresponds with the IARU-recommended S9 level at HF bands. By leaving the output unterminated, the open-circuit output voltage produced by every multiple of 100 kHz (*100 kHz* position) is 100 μV, or S9 at the high-impedance input terminal of an HF receiver. Terminated into a 50-ohm receiver antenna input, the voltage drops to the IARU-recommended 50 μV.

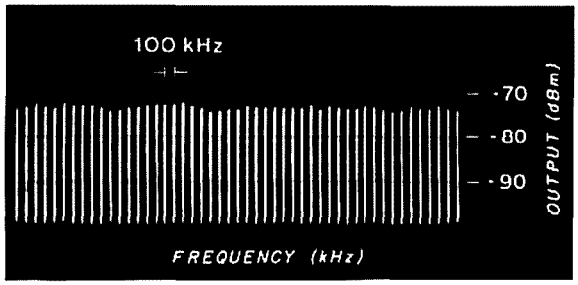
Once the receiver S-meter has been calibrated for this level, the other S-points are easy to determine by halving the voltage for each lower S-point on the scale. To put it more practically, each S-point change in level corresponds to 6 dB. Just a few examples: S8 = −79 dBm, S3 = −109 dBm.

Somewhere in the last NAND lies the borderline between the digital and analog sections of the circuit. From this point onward, carefree wiring concepts are better abandoned. The resistor network between the NAND and output connector requires careful layout consistent with good VHF practice. The resistors are best mounted free from each other and from other components, and with their wires as short as possible. Printed-circuit technique is not recommended here.

The shield box is electrically insulated from the circuit ground except at the output connector, through the coaxial cable.

oscillator

Experimenting with ICs of the 7400 series, one may find that the different blocks, with their low input and



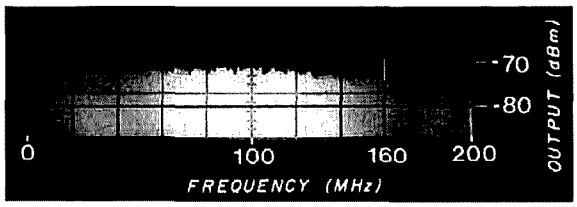


fig. 6. 100 kHz position. Top: typical segment of output frequency spectrum as displayed on spectrum analyzer. Each signal frequency is a multiple of 100 kHz and has a level of −73 dBm ±1.5 dB, up to at least 30 MHz. Bottom: spectrum displayed over full usable frequency range. Although signals are no longer individually distinguishable, the display shows signal levels within ±2 dB up to 160 MHz.

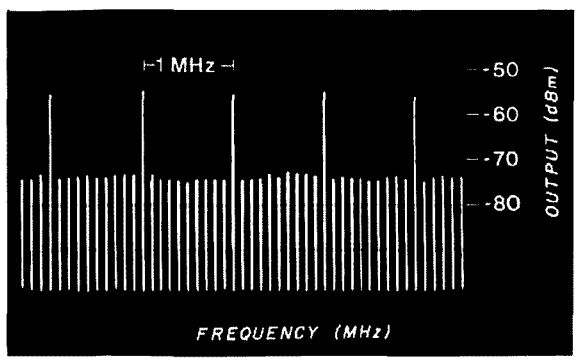


fig. 7. Position: 1 MHz + 100 kHz. Spectrum of 100-kHz marker signals, with 1-MHz marker amplitudes emphasized by approximately 20 dB.

output impedances, aren't the ideal elements for building a crystal oscillator. If the crystal isn't very active, the oscillator won't work. Nevertheless, the reason for using this circuit is that the NANDs — four of them combined in one IC — happened to be available, and they draw supply current whether they are used or not.

If the circuit doesn't work, it may well be possible to wake up a lethargic crystal by using different values for the resistors and capacitors in the oscillator circuit. However, it's also possible to overdo it, causing the oscillator to produce parasitics. In case of persistent recalcitrance, any other oscillator using one or two discrete transistors may do the job just as well, consuming only a few milliamperes of extra current.

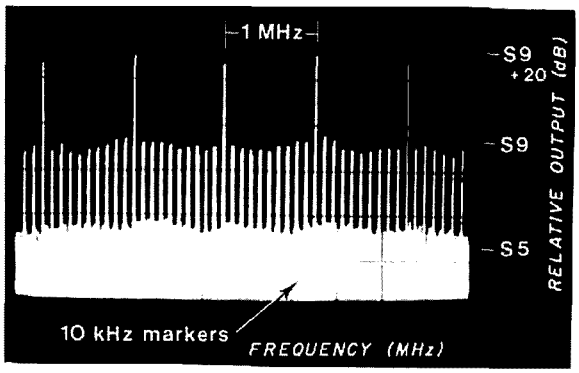


fig. 8. Position: 1 MHz + 100 kHz + 10 kHz. Spectrum is shown with all markers in operation.

power supply

One disadvantage of the closed-box technique may be that it doesn't leave much opportunity for monitoring the battery-supply voltage. That's why a voltage regulator, supplied by a battery that has some voltage to spare, is incorporated. The 9-volt battery can be drained to the point at which its actual voltage drops to 6.4 volts before the regulator gets in trouble trying to keep up its 5-volt output (that is to say, this is what I found; the 78L05 specifications state a minimum input voltage of 7 volts). As soon as the regulator output becomes affected, the RF output level of the total circuit becomes unreliable. Responsibility for this is mainly the last NAND in the output stage, producing pulses with lower amplitude and consequently less energy.

The problem of the aging battery is solved by applying a form of euthanasia. The "signal killer" circuit is based on the principle that it is better to have no signal at all than one on which you cannot rely. As soon as the battery drops below 6.4 volts, the RF output signal is cut off abruptly (**fig. 10**), before it can cause incorrect test results. This is done by blocking a NAND in the main signal path through a NAND inverter and a transistor.

When the appropriate moment arrives, the act is performed swiftly, even in case of a very slowly deteriorating battery voltage. The threshold is set by potentiometer R1 (**fig. 3**) to make the RF output signal disappear just before a reduction in amplitude occurs.

Because an output signal that suddenly disappears can be disconcerting, I found it useful to provide a means of alerting the user to the imminent demise of the battery. If the box fails to deliver its output signal, just turn it to the position *1 MHz + 100 kHz + 10 kHz*. If only the 10-kHz markers are still present, this is a sign that the malfunction of the generator is the result of too-low battery voltage, and that there's no reason to expect anything worse than having to replace the battery.

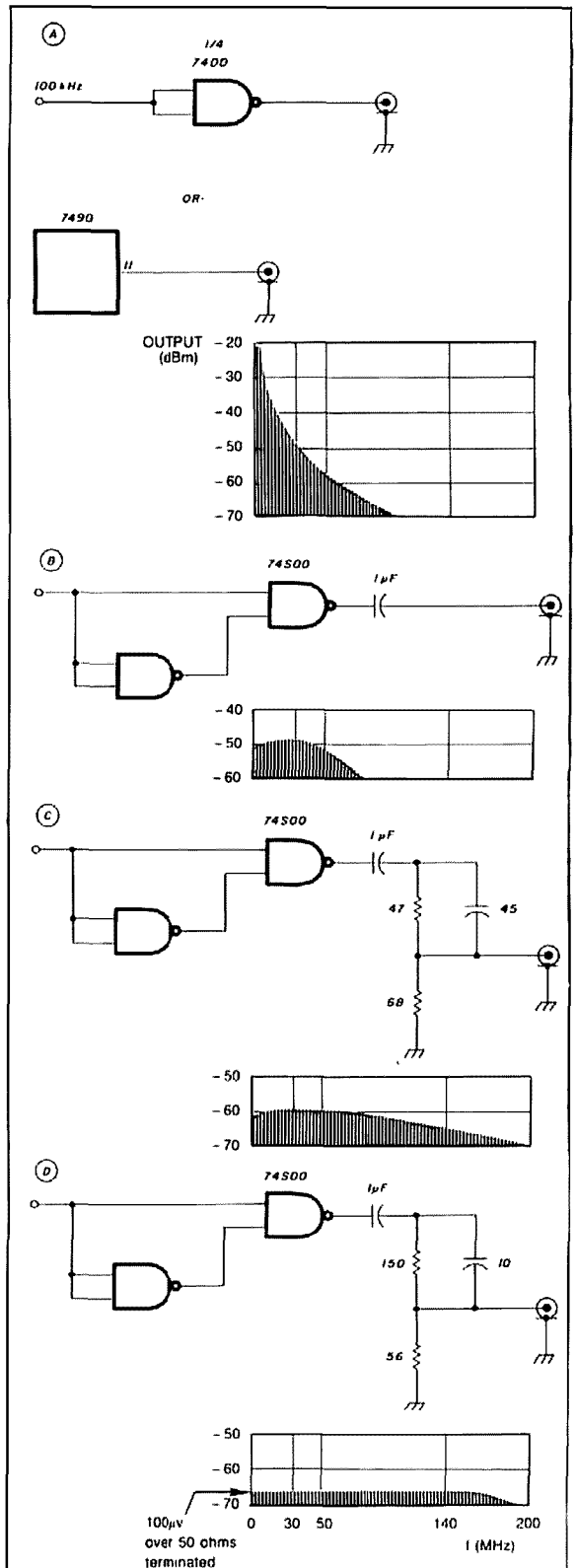


fig. 9. Examples of output-stage configurations and the effect on the frequency response. In all cases the output is terminated in 50 ohms.

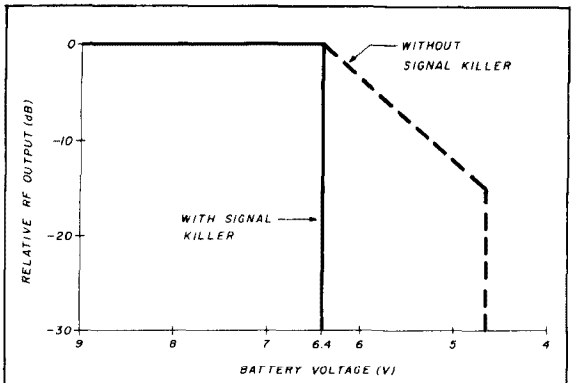


fig. 10. RF output as a function of battery voltage. An unnoticed drop in RF output level, resulting from an exhausted battery, is made impossible by the action of a "signal killer."

The battery is normally switched on and off by a mercury switch, a miniature glass bulb in which a droplet of mercury makes contact between two wires, depending on the position of the device. With some three-dimensional imagination, it's not difficult to see how, using a switch, a circuit can stay switched on at a maximum of three different positions of the box. This is done by mounting the switch in the circuit wiring at an oblique angle with respect to the three axes of the box. Similar switches are used for switching the 1-MHz and 10-kHz markers.

If mercury switches are difficult to obtain, it shouldn't be difficult to think of something else, such as one or more reed switches and a movable magnet, for example. Another method would be to use a steel ball (a ball bearing, for example) that can roll around inside a little plastic box with contact wires, or in a little box made of printed-circuit board (**fig. 11**).

The total battery consumption at 9 volts is about 43 mA with all markers operational, or about 36 mA in the *100 kHz* position.

attenuator

The type of outboard step attenuator shown in the photograph has been described already in many Amateur publications. In this case, miniature toggle switches and 1/4-watt carbon resistors are used. The resistors are mounted on the solder lugs of the switches, with their leads kept as short as possible. For the rest of the wiring, copper braid (taken from thin coaxial cable) is used, soldered over the full length of the solder lugs. Braid is not essential insofar that it doesn't seem to extend the flat portion of the frequency response; however, I found that it can give a considerably smoother result at VHF.

Production tolerances in the resistor-manufacturing industry appear to have greatly improved over the last

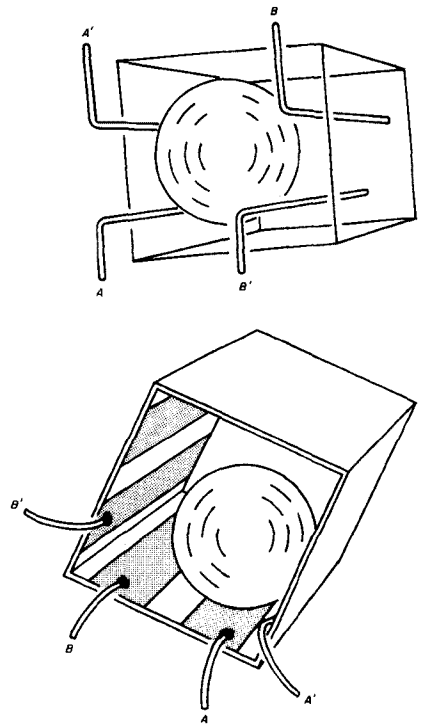


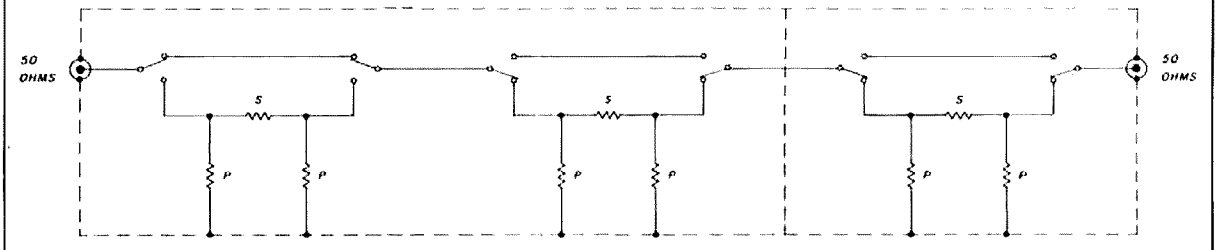
fig. 11. Substitutes for a mercury switch showing suggestions for an inclination switch using a steel ball. In position sketched: a-a′ closed contact, b-b′ open contact.

decade. At one time, a 10-percent resistor was almost bound to show a tolerance between 5 and 10 percent.

Today it's not unusual to find that ordinary 10-percent resistors are in fact not much more than 2 percent off the mark. And as more accurate ohmmeters become available to Radio Amateurs, it's no longer difficult to obtain reasonably precise resistors.

The required values for the attenuator are obtained by parallel combinations. In each combination it is mainly the lower-value resistor that is critical and which should preferably be within 1 to 2 percent. (Tolerances of resistors are like those of voltages, and the same rule-of-thumb applies: for small variations, each percent corresponds to 0.1 dB.) Alternatively, each combination should be brought as close as possible to the ideal value by selection and trial and error, using an accurate (digital) ohmmeter (**fig. 12**).

Most descriptions of this type of step attenuator remain rather vague about the performance, particularly at higher frequencies. Therefore, **fig. 13** may be of interest, because it shows the test results of this home made attenuator, built with no particular care other than ordinary construction techniques. The device is accurate within 1/10 of a dB or thereabouts,



	parallel resistor P ideal	parallel resistor P practical	series resistor S ideal	series resistor S practical
1 dB	869.5	1K//6.8K	5.765	6.8//39
2 dB	436.3	470//6.8K	11.62	12//390
3 dB	292.4	330//2.7K	17.62	18//820
6 dB	150.4	150	37.35	39//820
10 dB	96.25	100//2.7K	71.15	82//560
20 dB	61.10	68//560	247.5	270//2.7K

fig. 12. Design parameters of toggle-switch attenuator with standard-value resistors. Practical column lists parallel combinations that most nearly meet ideal values. (All values in ohms.) Note: // means placing the two resistors in parallel.

up to 65 MHz; that is, as accurate as the resistors could be selected with the available DC ohmmeter.

Figure 13 also shows the reason for the apparent reluctance to publish test results: at VHF precision tends to become questionable. Nevertheless, the general performance is still within the tolerance of the signal generator, about ±1 dB uncertainty at 144 MHz.

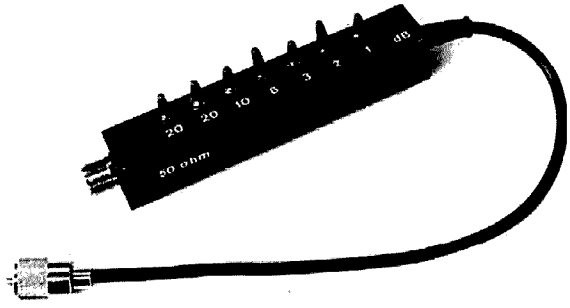
That the discrepancies are mainly the result of the many twists and turns in the signal path through the switches is also clearly demonstrated. Even with all attenuators in the *OUT* position (*0 dB*), the wavy character of the curve remains. This gives hope; it's not impossible that (cheaper) slide switches and N-type coaxial connectors would make an improvement on VHF.

calibration

The accuracy of the frequency can be checked and corrected by adjusting the oscillator trimmer while comparing a signal with one of the short-wave standard signals such as WWV in the USA, or GDF in Europe. Because the circuit doesn't contain any variables in the form of tuned circuits, the output level of −73 dBm should be fairly easily reproducible for HF by using the components as described. Up to, say 30 or 40 MHz, the frequency response should remain flat within reasonable limits, even if the output trimmer is simply replaced by a 10 pF fixed capacitor.

In the multitude of standard-size 100-kHz harmonics there appear to be two exceptions where the amplitude seems to behave a bit freakishly: at 100 kHz itself and at 1MHz. I found relative levels of −4 dB and −10 dB, respectively. As yet I have no fully satisfactory explanation to offer for the phenomenon; it's difficult to speculate to what extent these findings may be typical for the circuit.

The output can be calibrated by comparing the level



Toggle-switch attenuator.

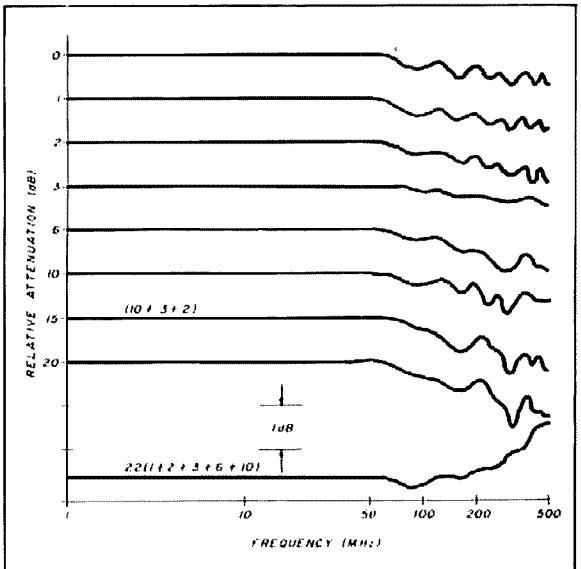
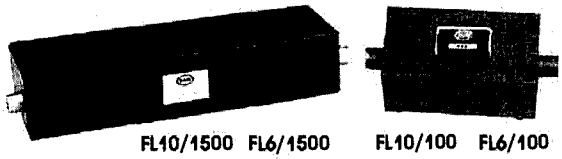


fig. 13. Test results of home made attenuator. Frequency response is shown at each setting, as well as at some combinations. Note absence of any appreciable discrepancies up to 65 MHz.

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of one of the 100-kHz harmonics (except at 100 kHz and at 1 MHz) with that of a signal from an RF generator with a reliable output attenuator, using a radio receiver as a frequency-selective balance detector. A signal-strength meter is not absolutely essential. By comparing two beat tones, one immediately after the other, providing that they both have the same pitch, one should be able to match levels well within 1 dB.

However, a professional RF signal generator could be difficult to procure, and for the average Radio Amateur a more common audio generator might be more easily available. Audio levels are more accurately determined using an audio voltmeter or an oscilloscope. Most audio generators cover 200 kHz, and a level comparison could be made on a long-wave radio receiver, using an audio meter or the human ear. After all, the 200-kHz harmonic is just as strong as any other harmonic in the entire HF spectrum of the generator and, even if it sits at an extreme end of the range, it provides a perfectly typical specimen upon which to base the level calibration.

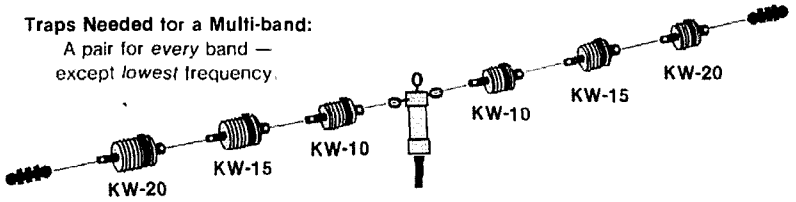
In case level corrections appear to be necessary, the resistors in the attenuator between output trimmer and coaxial cable can be replaced to change the 6-dB attenuation into a different value. **Figure 13** supplies the data.

ham radio

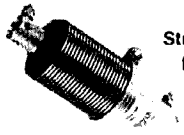
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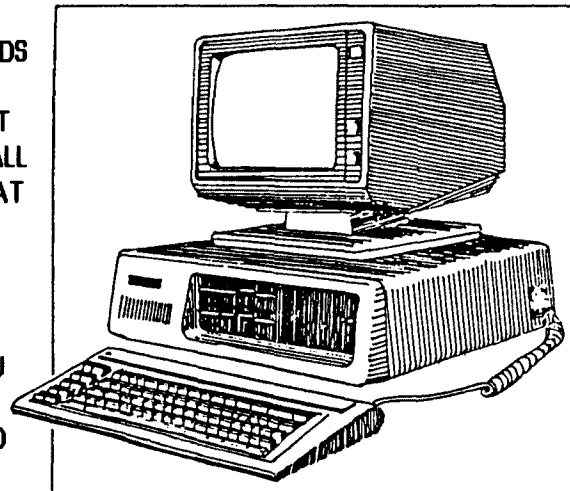
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More Details? CHECK—OFF Page 126

a calibrated S-meter

Meaningful signal reports *are* possible — here's how

Long ago a **wonderful idea emerged** that would make signal reporting, gain measurements, antenna comparisons, and propagation studies as easy as reading, reporting, and comparing meter values. This idea came to be known as the S-unit system of signal reporting. Sometime later an S-unit "standard" was set, in which each S-unit change represented a level change of 6dB and S9 was referenced to 50 microvolts rms in a 50 ohm system. This standard was straightforward and simple. S1 was 48 dB below 50 microvolts, near the practical lower limit of receiver sensitivity and noise floor. This standard offered the Amateur community a method of quantifying signal reporting and other useful measurements. **Tables 1** and **2** are included to illustrate the nearly 180 dB of dynamic range common in Amateur Radio equipment and show where the S-unit scale fits in at the threshold of receiver sensitivity.*

Nearly half a century has passed with very little conformation to this or any other level standard. For practical purposes, the RST system as practiced today is a farce — yet it's the most common exchange between stations. Consequently, meaningless guesstimates echo endlessly through the ether, apparently for lack of something more meaningful to say. This ironic state of affairs is especially troublesome in light of the advantages standardization and conformation could afford.

By nature, Amateur Radio is a quantitative hobby embracing the technology and science of multiple disciplines. Why, then, is the basic RST number abused? The more cynical among us might suggest that we're simply locked into a bad habit, or that you must give S9 signal reports to get rare QSL cards. There's probably some truth to the first notion; the second, even if true, is a poor reason for general abuse. After careful investigation, however, more subtle reasons emerge. The basic problem is simply that modern receivers generally do not have calibrated S-meters. Standards differ drastically from one manufacturer to another and from one unit to another. Most Amateurs know their S-meters are probably not calibrated and consequently give politely inflated signal reports.

Receiver manufacturers will respond only to customer demand; this is reflected by the multitude of "bells and whistles," compact size, and sleek styling found in modern Amateur equipment. Because Amateurs appear not to care, important specs have actually declined over the past several years. Dynamic range, on the other hand, has been improved because it has received much attention in the literature. Amateurs have become aware of this important phenomenon and now often ask equipment dealers embarrassing questions. I hope this signals a trend toward increased consumer demand for improved radio performance rather than for superficial enhancements that may have little or no effect on actual performance.

To be fair, lack of concern from the Amateur community is not the only reason that receiver manufacturers have not offered calibrated S-meters. There are real technical problems inherent in doing so. Providing consistent front end mixer gain over many frequency bands, for example, is a major problem. All amplifiers and attenuators ahead of the S-meter circuit must have identical gain on all bands and frequencies within these bands to assure consistent readings. Furthermore, most S-meter voltage sources are derived from the AGC circuit, and the response of these circuits varies widely — not only among receiver model types, but among individual units of similar types as

By Robert J. Zavrel, Jr., W7SX, P.O. Box 24845, San Jose, California 95154

* From minimum discernible signal at the receiver to maximum legal output from a linear — **Ed.**

well. **Table 3** shows the extent of this deviation among popular receiver types. [1,2,3] This unpredictable deviation from receiver to receiver of the AGC voltage versus received signal strength represents the most difficult obstacle to setting an S-meter standard. The first problem can now be solved by broadbanding front ends, using flat response mixers, and employing modern RF components and layout techniques. The second problem has been alleviated with the introduction of the Signetics NE604, an FM RF integrated circuit designed primarily for the cellular radio market (**fig. 1** and **2**).

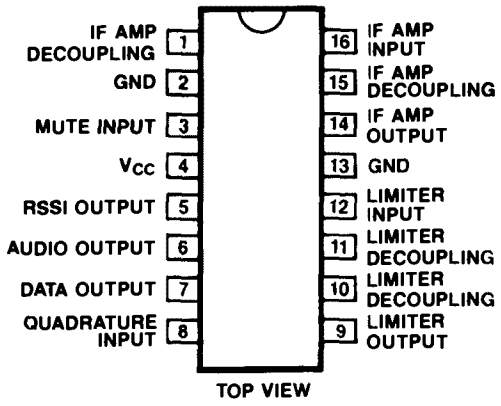


fig. 1. Pin-out of NE604.

Receivers for cellular radio systems require a received signal strength indicator (RSSI) function. The response of this required logarithmic measurement must conform to 3 dB accuracy over an 80 dB dynamic range. This function represents an ideal signal source for S-meters and other RF measurement equipment

table 1. Power and voltage levels common to Amateur equipment.

dBM	power (watts)	volts (50 ohms)
+70	10 kW	707.0 V
+60	1 kW	223.0 V
+50	100 W	70.7 V
+40	10 W	22.3 V
+30	1 W	7.07 V
+20	100 mW	2.23 V
+10	10 mW	707.0 mV
0	1 mW	223.0 mV
−10	100 μW	70.7 mV
−20	10 μW	22.3 mV
−30	1 μW	7.07 mV
−40	100 nW	2.23 mV
−50	10 nW	707.0 μV
−60	1 nW	223.0 μV
−70	100 pW	70.7 μV
−80	10 pW	22.3 μV
−90	1 pW	7.07 μV
−100	100 fW	2.23 μV
−110	10 fW	707.0 nV
−120	1 fW	223.0 nV

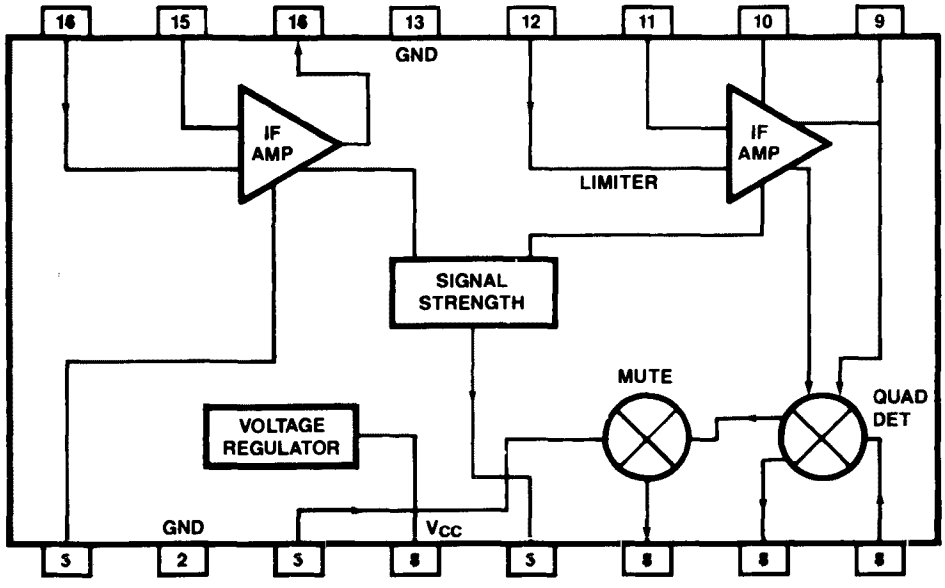


fig. 2. Functional diagram of NE604.

table 2. Power and voltage levels for the nine S-units.

S-unit	dBm	volts (50 ohms)
S9	−73	50.0 μV
S8	−79	25.0 μV
S7	−85	12.6 μV
S6	−91	6.3 μV
S5	−97	3.2 μV
S4	−103	1.6 μV
S3	−109	792.0 nV
S2	−115	397.0 nV
S1	−121	199.0 nV

such as field strength meters. The NE604 performs this function by providing a current source which is typically sunk through a 100 kilohm resistor (**fig. 3**). The high impedance voltage produced at pin 5 is then buffered by an op amp (NE532) to drive a linear analog voltmeter or other suitable display. The DC voltage output is plotted against RF level input in **fig. 4**. Responses for unattenuated and attenuated inputs are given.

The RF input to the NE604 can be taken off an IF amplifier stage operating up to 15 MHz. The higher RF levels associated with the later IF stages make them more attractive as drivers for the 604. Very light coupling can be employed (3 pF) with these stages to prevent any undesirable loading to the receiver IF amplifier. **Figure 4** shows that the unattenuated amplitude input range for the 604 is about −90 to −10 dBm. The effect of the **fig. 3** input isolation circuit is to move the effective range up about 20 dB. Because the RSSI function is mode independent, the S-meter will function for CW, AM, SSB, or FM.

A simple LC filter should be placed between the IF and limiter stages (pins 12 and 14) to reduce broadband noise. Optimum performance of the 604 occurs with a 6-volt power supply, but will work over a 4 to 8 volt range. Supply current is only about 2.5 mA suggesting battery powered operation for both Amateur receivers and accurate field strength meters.

building an S-meter circuit

As stated earlier, each S-unit represents a 6 dB change in RF signal strength. For each 10 dB change in RF input, the 604 circuit output will change about 1/2 volt. If S1 is referenced to the lower limit of the 604's dynamic range (about 0.3 volts output), then S9 will be 48 dB higher (about 2.7 volts output.) Each S-unit will represent 0.3 volt of output meter calibration. A standard linear 0-5 volt meter can be calibrated for S-units and "dB over S9." With an 80 dB dynamic range and 48 dB difference between S1 and S9, 32 dB remains for the "dB over S9" calibration. Furthermore, the S-meter can also be calibrated in dBm and/or RF

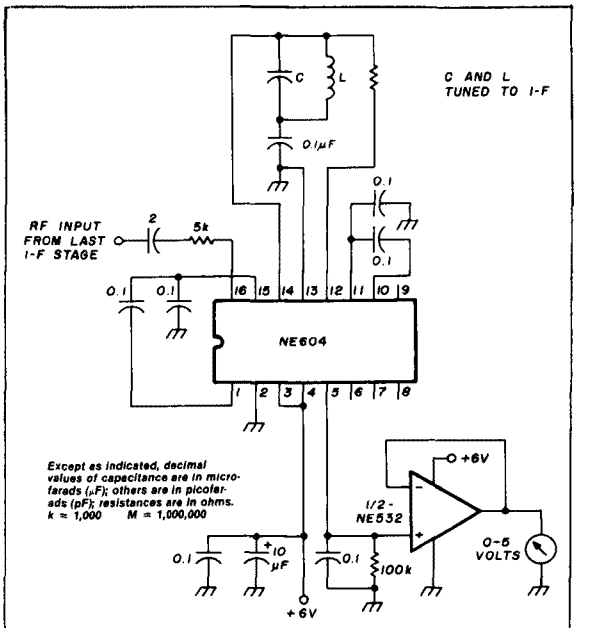


fig. 3. Complete S-meter circuit.

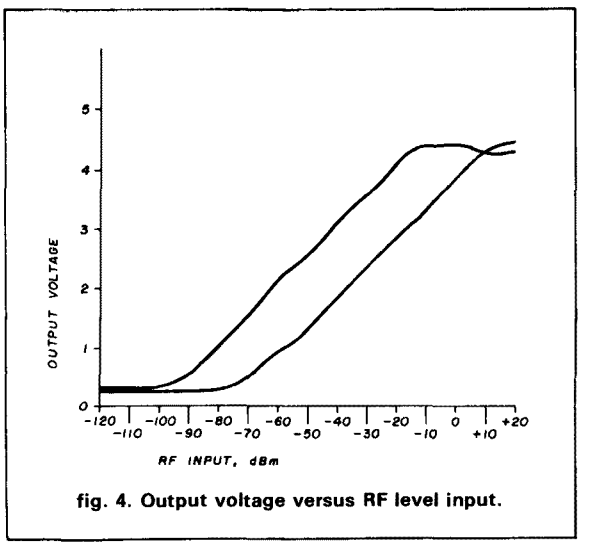


fig. 4. Output voltage versus RF level input.

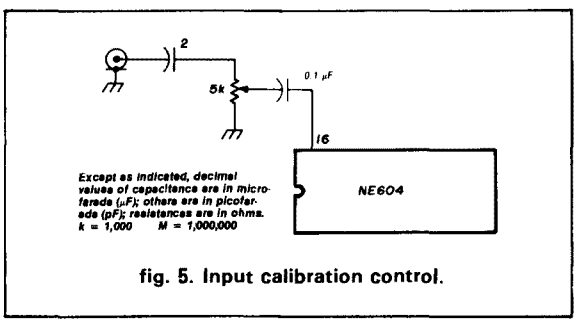


fig. 5. Input calibration control.

volts. The receiver has now evolved into a valuable RF test instrument.

Because it might overshoot the actual peaks of signal intensity, an underdamped meter is not recommended for S-meters; critically or slightly overdamped meter movement is preferred instead. An LED or LCD bar graph would also make a suitable display. Digital readout displays are not recommended for S-meters because of their inferior peak reading response.

The RF input impedance at pin 16 is about 1500 ohms. Other capacitor and resistor values can be set for a given IF signal level range, including a potentiometer for a calibration adjustment (**fig. 5**). Calibration consists of placing a 50 microvolt signal at the antenna terminal, peaking the response by fine tuning the VFO, and adjusting the pot for S9. Correct meter tracking will now be automatic.

This circuit will provide an accurate S-meter reading if certain precautions are taken. Most S-meters use the AGC voltage as the control voltage. Although the RSSI generated voltage would also be ideal for the AGC voltage source, the S-meter linearity would require the AGC function to be disabled. Alternatively, the meter can be calibrated with the AGC "on". Either way, the NE604 will also allow you to know exactly the amount of gain limiting present at any time by comparing the S-meter readings with and without AGC. When not using the AGC, any RF or IF gain controls must have a "calibrated" position. Furthermore, if gain varies with frequency, as in older receiver designs, any array of trim pots controlled by the bandswitch could be the answer.

An important feature of this circuit is that relative dB measurements will be very accurate even if the absolute reading is not calibrated. This is a useful practical on-the-air feature for accurate antenna gain comparisons.

Figure 4 shows that the output voltage never goes to zero. A meter zeroing control should be easy to build with the addition of a few more op amps and standard DC offset techniques (**fig. 6**).

This is a simple project that can transform your receiver into a field strength and RF voltmeter. More importantly, it can transform meaningless numbers into valuable measurements.

table 3. Tested S9 μV sensitivity for some popular rigs.

radio	S9 (μV)	radio	S9 (μV)
FT901	50	Omni D	36
TS-820	200	Triton IV	20
SB104	265	Atlas 350 XL	150
FT301	95	FT901	8
TS-520	110	Argonaut	10
FT7	180	TS-820	110
TS-120V	180	FT301S	30
TR7	20	SB303	70
75S3B	250	KWM2	60
IC701	20	FT101E	10

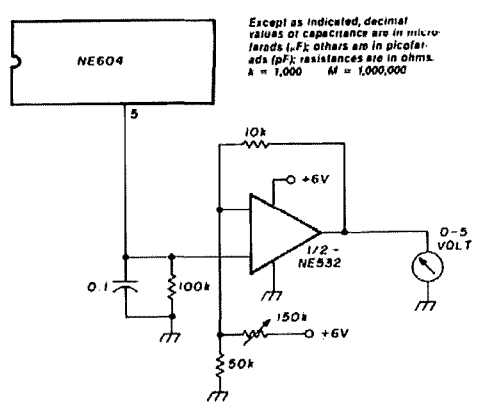


fig. 6. Meter zero calibration.

references

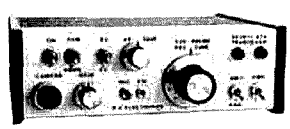
1. *CQDL*, December, 1980, pages 550-555.
2. Sidney Kaiser, WB6CTW, "Measuring Receiver Dynamic Range," *ham radio*, November, 1979, page 56.
3. Doug DeMaw, W1FB, and Wes Hayward, W7ZOI, "Modern Receivers and Transceivers: What Ails Them?" *QST*, January, 1983, page 11.

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passive audio filter design

part 3: attenuation equalizers and driving, terminating audio filters

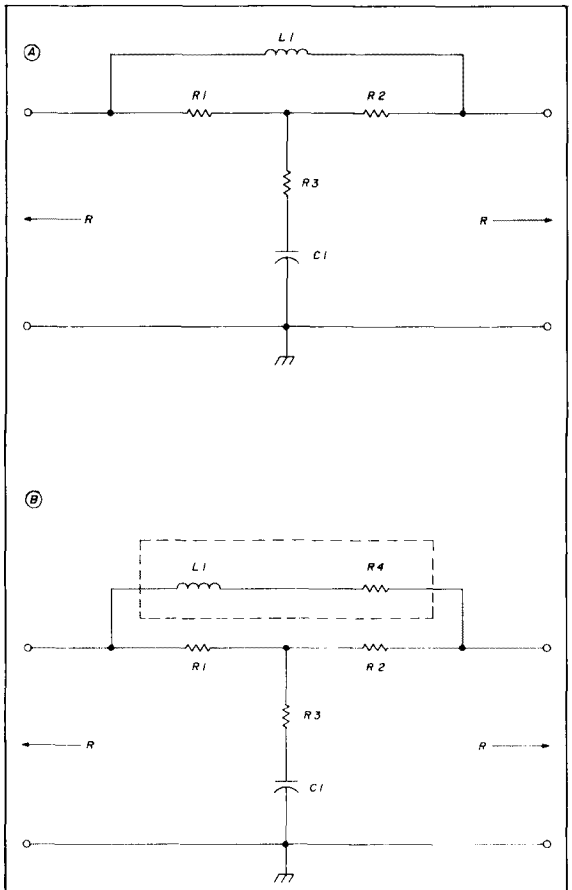


fig. 1. Low-pass attenuation equalizer.
(*A*) schematic of equalizer with ideal inductor;
(*B*) schematic of equalizer with real inductor.

Improving the response of filters constructed with low-Q inductors

In parts 1 and 2[1,2] of this series, it was shown that two effects of using low-Q inductors in a filter are a general decrease of the passband response as frequency increases and a roundoff near the theoretical cutoff frequency. These effects are most noticeable with the responses of sharp cutoff filters, such as the elliptic low-pass filters described earlier. To a certain extent, these effects can be reduced by adding a high-pass filter in series with the low-pass, giving a band-pass response.

The classical method of correcting this roundoff of the amplitude response (other than by using higher-Q inductors) is by using an attenuation equalizer. Do not be put off by the name; equalizers are very simple and can produce a dramatic improvement in filters with low-Q inductors. I will now outline the design of these equalizers and show the effect they can produce on some of the filters already described.

low-pass attenuation equalizers

The schematic of an ideal low-pass attenuation equalizer is shown in **fig. 1(A)**, along with equations

By Stefan Niewiadomski, 29 Mackinley Avenue, Stapleford, Nottingham, NG9 8HU, England

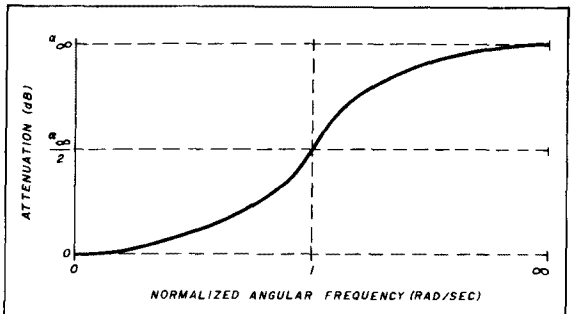


fig. 2. Response of ideal low-pass attenuation equalizer.

that define the normalized (1-ohm, 1-rad/sec) component values in terms of a constant, K.[3] **Figure 1(B)** shows the equalizer with a resistor, R4, added in series with L1 to simulate the low Q of the inductor used in the actual equalizer. The theoretical response of the ideal network is shown in **fig. 2.**

At zero frequency, the equalizer has no loss. As frequency increases, the attenuation increases, asymptotically approaching a finite value of α_∞ dB at infinite frequency. The value of α_∞ is given by:

$$\alpha_\infty = 20 \log K$$

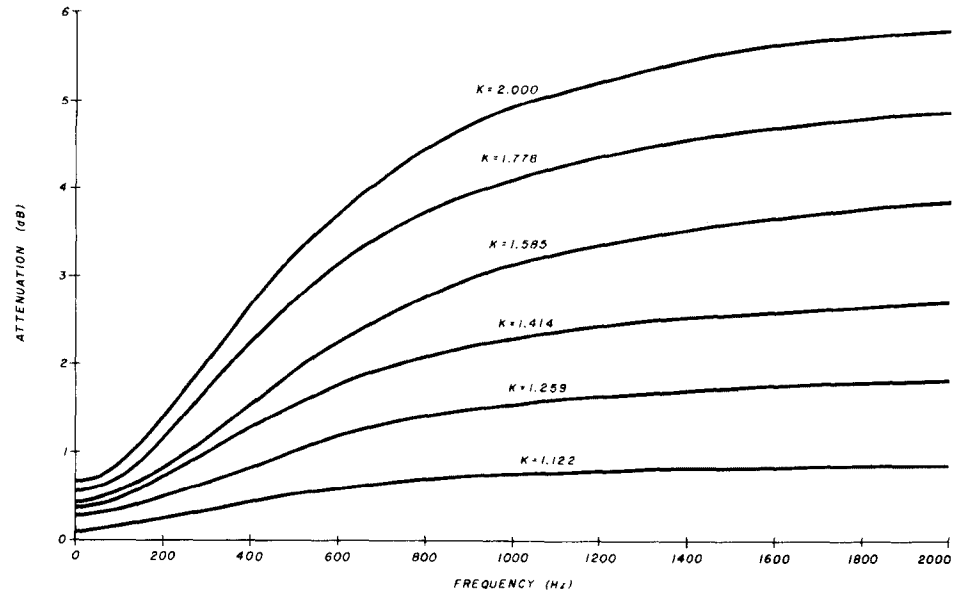


fig. 3. 0-2 kHz response of 500-ohm, 500-Hz low-pass attenuation equalizer showing simulation results for various values of K.

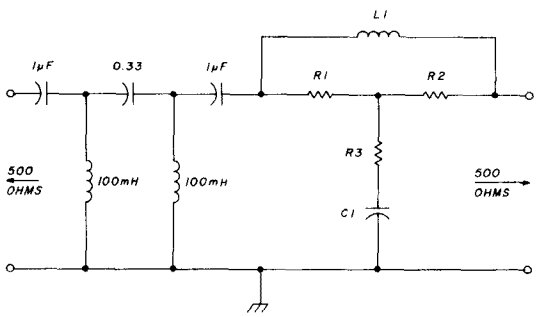


fig. 4. Schematic of 500-ohm, 500-Hz Butterworth highpass filter in series with 500-ohm, 500-Hz low-pass attenuation equalizer (values of equalizer components given in *table 1*).

where K is the constant used to calculate the component values. At the normalized angular frequency of 1 rad/sec, the attenuation is $\alpha_\infty/2$.

A low-pass equalizer is intended for use with a highpass filter, constructed with low-Q inductors, to correct its response near its cutoff frequency. I will use the Butterworth 500-ohm, 500-Hz highpass filter described previously to illustrate the effects of a low-pass equalizer. The normalized equalizer component values must therefore be scaled to 500 ohms to match the highpass filter. The scaled equalizer will then present an impedance at its terminals which is purely resistive and equal to 500 ohms at all frequencies. The half-attenuation frequency of the equalizer does not necessarily have to be equal to the filter cutoff frequency, but this represents a good starting point. The final,

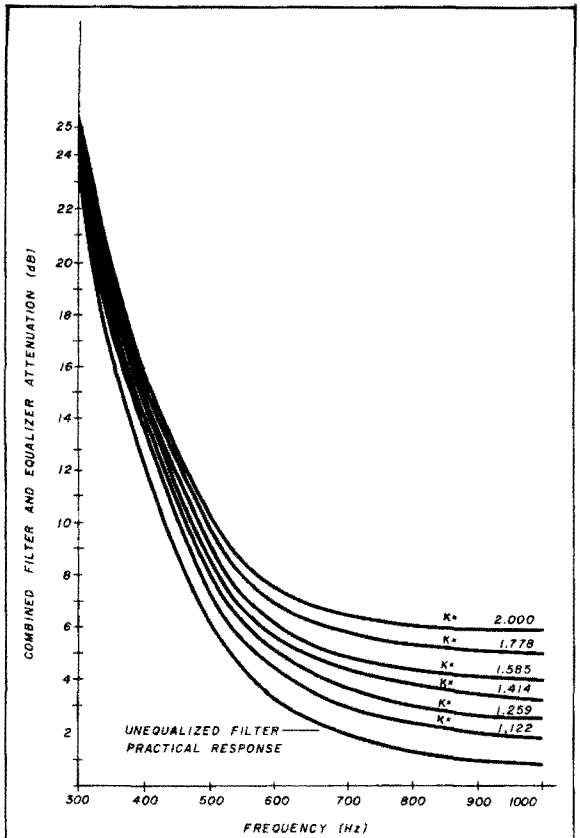


fig. 5. 300-1000 Hz response of 500-Hz Butterworth highpass filter with 500-ohm, 500-Hz lowpass equalizer for various values of K, showing simulation and practical results (original highpass filter practical response also shown for comparison).

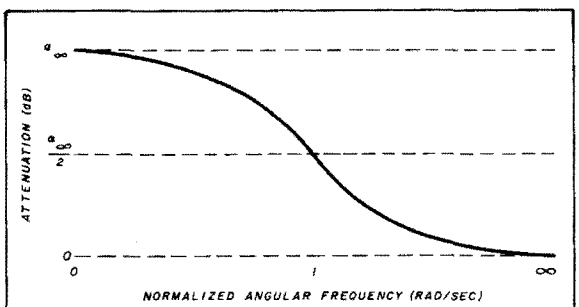


fig. 7. Response of ideal highpass attenuation equalizer.

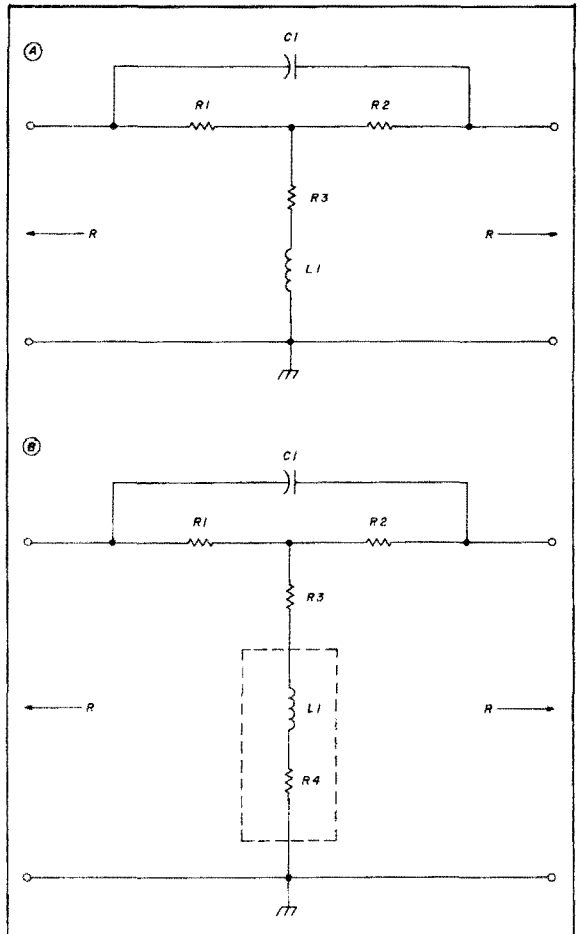


fig. 6. Highpass attenuation equalizer.
(A) schematic of equalizer with ideal inductor;
(B) schematic of equlaizer with real inductor.

scaled values of the equalizer are therefore set for 500 ohms and 500 Hz.

Table 1 shows, for six values of K, the 1-ohm, 1-rad/sec component values; the exact 500-ohm, 500-Hz values assuming an ideal inductor; and rounded values with a value entered for R4 to model the real inductor, L1. These six K's give values of α_∞ of 1,2, 3, 4, 5, and 6 dB, respectively, and attenuations at 500 Hz (for the scaled values) of 0.5, 1, 1.5, 2, 2.5, and 3 dB. Component values for the scaled impedance and half-attenuation frequency are obtained in the same way as for filters,* so the normalized values in **table 1** can be used to obtain practical values for any impedance and half-attenuation frequency.

Figure 3 shows the simulated response of the

*See part 1, September, 1985, page 30.

500-ohm, 500-Hz equalizer with the rounded values and real inductor. Curves are plotted for the six values of K tabulated in table 1. For K = 1.122, 1.259, 1.414, and 1.585, good agreement with theory was found, the half-attenuation values being 0.5, 1, 1.55, and 1.95 dB, respectively. At the maximum frequency plotted, 2 kHz, the responses flatten out and approach the maximum attenuation of 1, 2, 3, and 4 dB. The curves for K = 1.778 and 2.000 show more deviation from the theoretical responses, having half-attenuation values of 2.7 and 3.25 dB. However, at 2 kHz the maximum attenuations of 5 and 6 dB are approached.

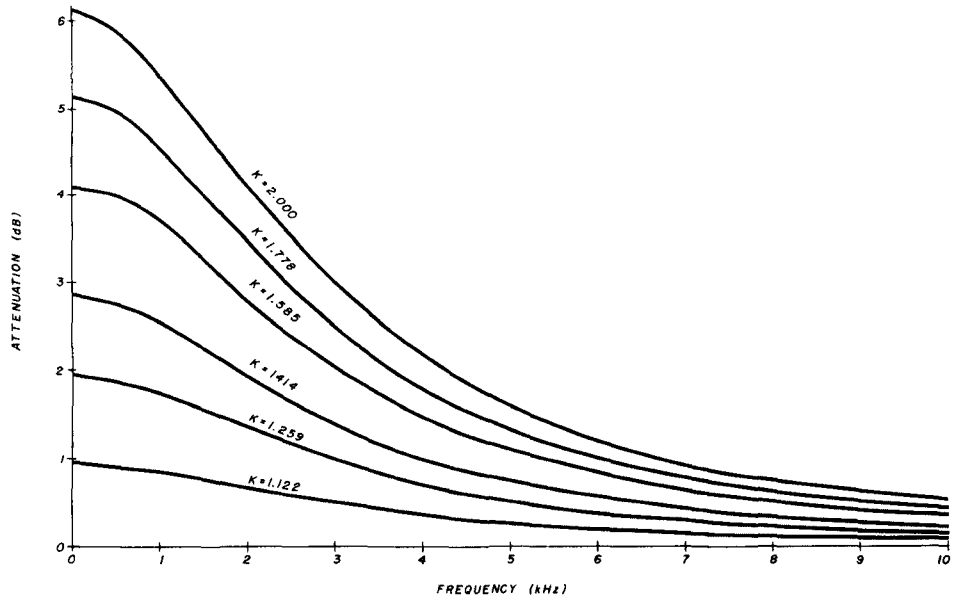


fig. 8. 0-10 kHz response of 500-ohm, 3-kHz highpass attenuation equalizer showing simulation results for various values of K.

table 1. Low-pass attenuation equalizer component values.

component	1 ohm, 1 rad/sec value	500 ohm, 3 kHz theoretical value ideal inductor	500 ohm, 3 kHz rounded value real inductor	1 ohm, 1 rad/sec value	500 ohm, 3 kHz theoretical value ideal inductor	500 ohm, 3 kHz rounded value real inductor
	K = 1.122				K = 1.259	
C1	0.1152 F	0.0733	0.068	0.2308 F	0.1469	0.15
L1	0.1152 H	18.33 mH	18 mH	0.2308 H	36.73 mH	39 mH
R1,R2	0.0575	28.75	27	0.1147	57.35	56
R3	8.668	4334	4700	4.304	2152	2200
R4	—	0	17	—	0	45
	K = 1.414				K = 1.585	
C1	0.3482 F	0.2217	0.22	0.4647 F	0.2958	0.27
L1	0.3482 H	55.42 mH	56 mH	0.4647 H	73.96 mH	68 mH
R1,R2	0.1715	85.75	82	0.2263	113.15	120
R3	2.830	1415	1500	2.0962	1048	1000
R4	—	0	58	—	0	66
	K = 1.778				K = 2.000	
C1	0.5835 F	0.3715	0.39	0.7071 F	0.4502	0.47
L1	0.5835 H	92.87 mH	100 mH	0.7071 H	112.54 mH	120 mH
R1,R2	0.2801	140.1	150	0.3333	166.67	180
R3	1.6453	822.7	820	1.3333	666.67	680
R4	—	0	82	—	0	97

Having shown by simulation that low-pass equalizers can be built with practical, preferred-value components, the effect on the Butterworth 500-ohm, 500 Hz filter will now be shown.

The schematic of the Butterworth filter combined with the low-pass equalizer is shown in **fig. 4**. **Figure 5** shows the simulated response of the Butterworth highpass filter in series with the 500-ohm, 500 Hz equalizer. Again, the responses for six values of K are shown, as well as the practical response of the original filter. How much improvement in the response has occurred?

If we take the difference in attenuation between 1 kHz and 500 Hz (which should, in theory, be very close to 3 dB) as a measure of quality, the improvement can be assessed. For the unequalized filter, this figure is about 5.5 dB; for the filter with the equalizer of K = 1.122 it is about 5.7 dB; with K = 1.414, it is about 5.3 dB; and with K = 2.000, it is about 4.3 dB. There has been a gradual improvement in the response as K increases, but of course the price to be paid is a corresponding increase in the 1-kHz (and beyond) insertion loss. This loss can be compensated by increasing the gain of the active devices in the audio path. With this particular example, however, it is doubtful whether the fairly marginal improvement in response is worth the extra components required and the extra insertion loss suffered.

highpass attenuation equalizers

The schematic of an ideal highpass attenuation equalizer is shown in fig. **6A** with equations for the normalized component values. This equalizer has been obtained by an identical transformation to that applied to a lowpass filter to transform it into a highpass filter.

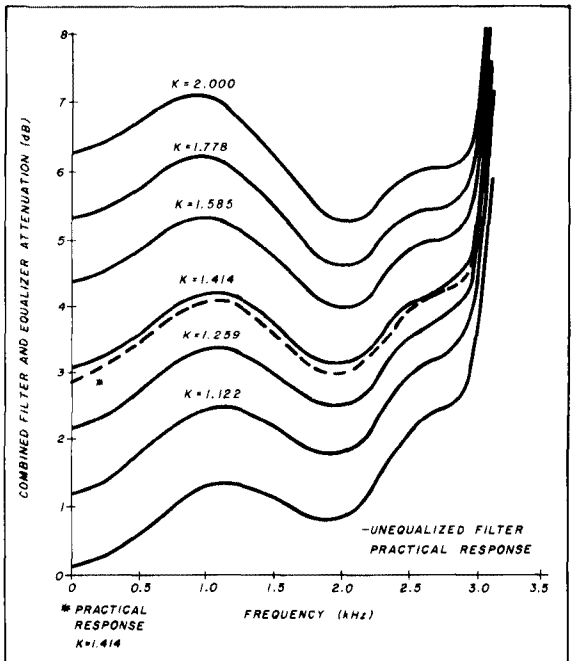


fig. 9. 0-3 kHz response of 1-dB/50-dB 5-branch 3-kHz elliptic low-pass filter with 500-ohm, 3-kHz highpass equalizer for various values of K, showing a simulation and practical results (original low-pass filter practical response also shown for comparison).

table 2. Highpass attenuation equalizer component values.

component	1 ohm, 1 rad/sec value	500 ohm, 3 kHz theoretical value ideal inductor	500 ohm, 3 kHz rounded value real inductor	1 ohm, 1 rad/sec value	500 ohm, 3 kHz theoretical value ideal inductor	500 ohm, 3 kHz rounded value real inductor
	K = 1.122				**K = 1.259**	
C1	8.681 F	0.921	1μF	4.333 F	0.460	0.47
L1	8.681 H	220.3 mH	220 mH	4.333 H	114.95 mH	120 mH
R1,R2	0.0575	28.75	27	0.1147	57.35	56
R3	8.668	4334	4700	4.304	2152	2200
R4	—	0	95	—	0	97
	K = 1.414				**K = 1.585**	
C1	2.872 F	0.3047	0.33	2.152 F	0.2283	0.22
L1	2.872 H	76.19	82 mH	2.152 H	57.09 mH	56
R1,R2	0.1715	85.75	82	0.2263	113.15	120
R3	2.830	1415	1500	2.0962	1048	1000
R4	—	0	71	—	0	58
	K = 1.778				**K = 2.000**	
C1	1.714 F	0.1819	0.18	1.4142 F	0.1500	0.15
L1	1.714 H	45.47 mH	47 mH	1.4142 H	37.52 mH	39 mH
R1,R2	0.2801	140.1	150	0.3333	166.67	180
R3	1.6453	822.7	820	1.3333	666.67	680
R4	—	0	52	—	0	45

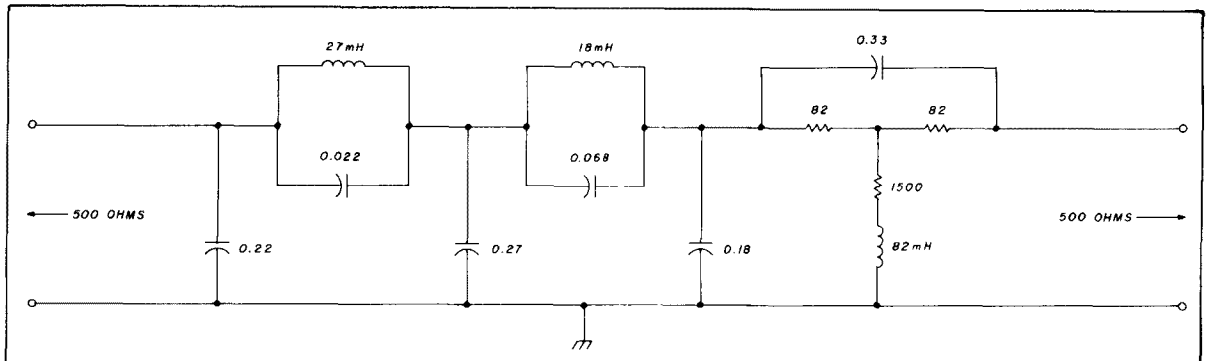


fig. 10. Schematic of 1-dB/50-dB elliptic low-pass filter in series with highpass equalizer for K = 1.414.

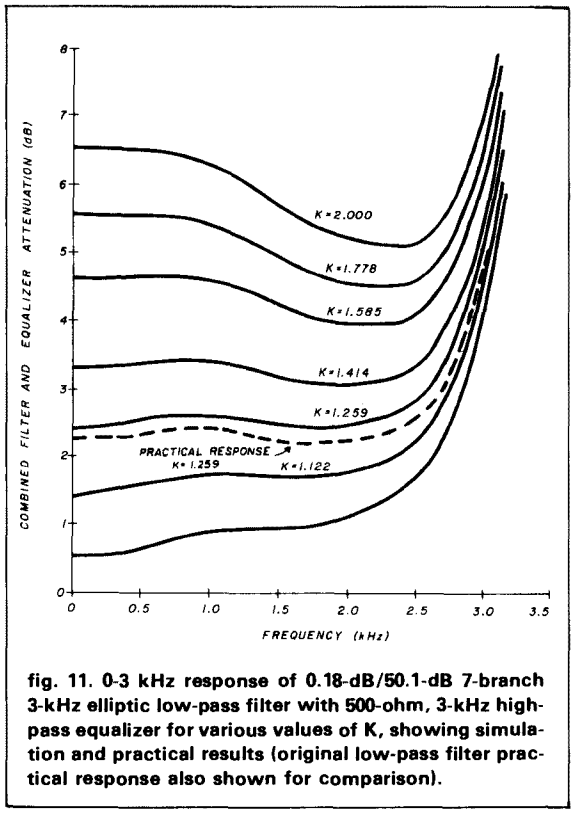


fig. 11. 0-3 kHz response of 0.18-dB/50.1-dB 7-branch 3-kHz elliptic low-pass filter with 500-ohm, 3-kHz highpass equalizer for various values of K, showing simulation and practical results (original low-pass filter practical response also shown for comparison).

That is, the inductor has been transformed into a capacitor, the capacitor into an inductor, and their values are the inverse of the original values. **Figure 6B** shows resistor R4 added in series with L1, as before. The theoretical response of the ideal network is shown in **fig. 7**. This response is the inverse of that of the low-pass equalizer.

The method of calculating $\alpha\infty$ is identical to the low-pass case, and again the half-attenuation angular frequency is 1 rad/sec.

Table 2 shows the component values obtained for the same six values of K as before. This time, however, the values have been scaled to 500 ohms, 3 kHz to make the equalizers compatible with the 1-dB/50-dB and 0.18-dB/50.1-dB low-pass elliptic filters described previously.

Figure 8 shows the simulated response of the 500-ohm, 3-kHz equalizer with the rounded values and real inductor. Curves for the six values of K are plotted. The simulations show good agreement with theory, having half-attenuation values close to the theoretical values at 3 kHz. For K = 1.122, 1.259, and 1.414, the curves approach 1, 2, and 3 dB respectively, as expected. However, for K = 1.585, 1.778, and 2.000, the asymptotic values are a little above those expected.

Figure 9 shows the simulated response of the 1-dB/50-dB elliptic low-pass filter in series with the highpass equalizers of table 2 for the six values of K. For these simulations, the equalizer was inserted after the filter, but which comes first is not important. The practical response of the filter with no equalization is also shown, and the progressive effect of increasing the value of K for the equalizer can be seen. Only the responses up to 3100 Hz are shown, because above this frequency they are almost indistinguishable. The difference in minimum stopband attenuation between the unequalized filter and the filter with the K = 2.000 equalizer is only about 1 dB.

As shown in **figure 9**, the smoothest passband response is obtained for K = 1.414. The passband ripple up to 260 Hz now meets the design specification of 1 dB, with approximately 1.5 dB more loss before the design cutoff frequency of 3 kHz is reached. The total attenuation difference between 100 Hz and 3 kHz is therefore approximately 2.5 dB, compared with approximately 3.6 dB for the unequalized filter. The penalty paid is a decrease in the differential between the minimum passband and minimum stopband attenuations from approximately 53 dB to approximately 50 dB.

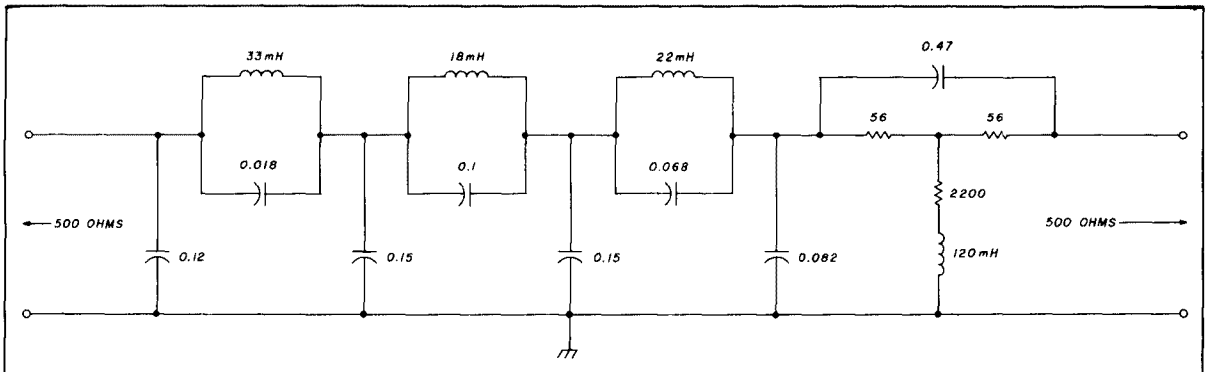


fig. 12. Schematic of 0.18-dB/50.1-dB elliptic lowpass filter in series with highpass equalizer for K = 1.259.

To verify the simulation results, the equalizer with K = 1.414 was built and connected to the output of the practical 1-dB/50-dB filter. The schematic of the combined network is shown in **fig. 10.** The results are plotted in **fig. 9** for comparison with the simulated response. Very close agreement with the simulated response was obtained. In practice, slightly less insertion loss is seen at all frequencies than in the simulated response.

The way in which the highpass equalizer modifies the 0.18-dB/50.1-dB elliptic low-pass filter is shown in **fig. 11.** The equalizer component values are those shown in table 2. A flattening of the passband response can be seen, particularly for K = 1.259. The equalizer for this value of K was built, placed in series with the 0.18-dB/50.1-dB filter and the combined response measured. The schematic of the combined networks is shown in **fig. 12**, and the practical results are plotted in **fig. 11** for comparison with the simulated response. Again, good agreement with the simulated response was found. The real insertion loss is slightly less than expected at all frequencies.

Ripple in the practical response for the combination with K = 1.259 is now less than 0.4 dB up to 2.5 kHz. Although this is still more than the theoretical maximum for the ideal low-pass filter, it is a considerable improvement on the unequalized response (approximately 1.5 dB).

driving and terminating audio filters

Whether passive audio filters are incorporated into existing equipment or included in new designs, correct rect source and termination impedances must be used if predictable results are to be obtained. **Figure 13** shows some methods commonly used to interface these filters.

Figure 13A shows a common emitter amplifier that uses Q1, which can be almost any commonly available transistor. The output impedance of this amplifier is equal to the collector resistor, R_S, so this should be chosen to be the same as the source impedance of the filter. The base and emitter resistors should be chosen to set the DC voltage at the collector of Q1 at a value that allows an AC swing compatible with the signal level being handled. The output buffer amplifier, Q2, is an emitter follower. The input impedance of the transistor in this configuration is very high, so the termination seen by the filter is the parallel combination of R1 and R2. R1 and R2 should therefore be chosen to terminate the filter correctly and set the DC operating conditions of Q2. If a DC path exists from the input of the filter to ground or to the output, then the input DC blocking capacitor, C_I, will be required. If a DC path exists from the output of the filter to ground, then C_O will be required. The RC network in the supply line prevents any signal coupling from input to output through the buss, which would degrade filter performance.

Transistor Q2 can be replaced by an FET stage. The termination impedance for the filter will then be provided by a resistor from the gate of the FET to ground.

Figure 13B shows an operational amplifier circuit. Because the output impedance of an operational amplifier with feedback is very low, the source impedance for the filter has to be provided by resistor R_S. The termination for the filter is provided by R_T. The feedback resistors for the operational amplifiers can be chosen to provide any reasonable value of gain for these stages. Of course, the techniques of **figs. 13A** and **13B** can be combined, and the final configuration can be chosen to make the best use of available components.

You may have noticed that the source and termination impedances for all the filters previously described is 500 ohms, which, of course, is a nonpreferred value. The importance of providing the correct impedances is often stressed, so how critical is

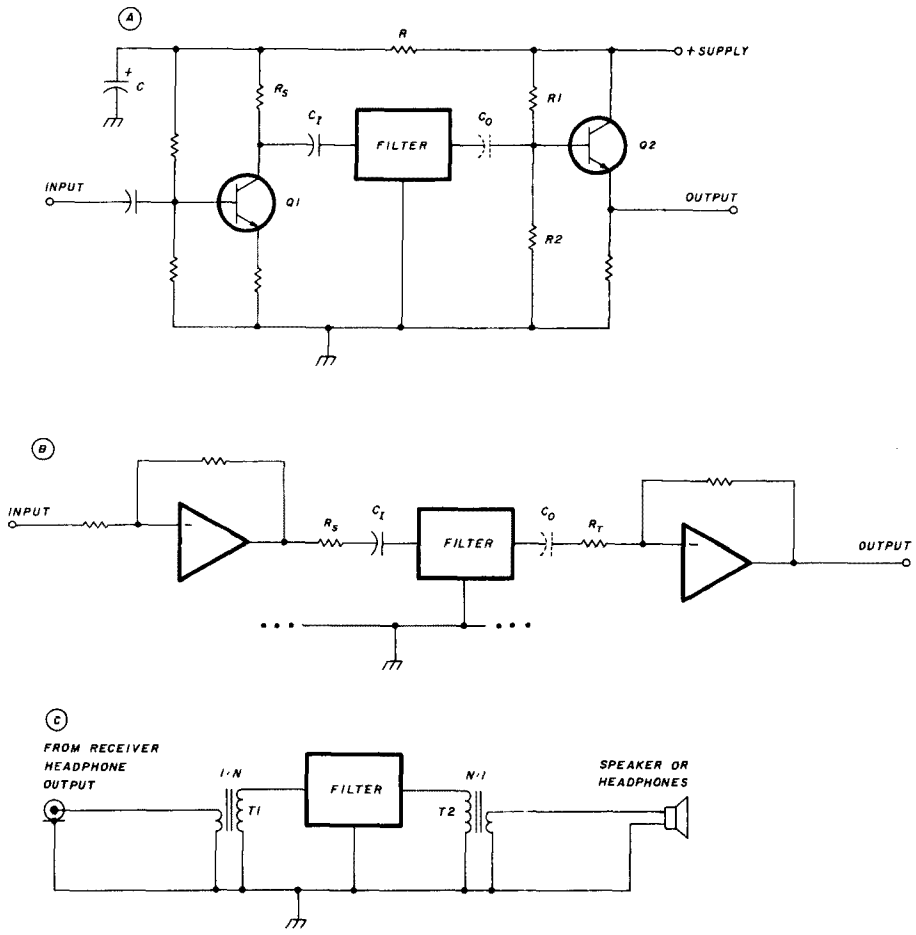


fig. 13. Various methods of connecting an audio filter into the audio path: (*A*) transistor circuit; (*B*) op-amp circuit; and (*C*) transformer-coupled circuit.

this matching? My own experiments have shown that the E12 values* of 470 and 560 will give slightly increased passband ripple, but not enough to be noticeable in practice. One interesting point is that an increase in passband ripple, even if it is caused by mismatching a filter, tends to give an increase in stopband attenuation (though this is not recommended as a method of obtaining better stopband performance.)

Often the simplest way of adding an audio filter is in the low-impedance headphone output of a receiver. A method used to couple the comparatively high-impedance filter (typically 500 ohms) to the much lower impedance audio output is shown in **fig. 13C**. Transformers T1 and T2 have a turns ratio equal to the square root of the ratio of the filter and speaker or phone impedances. One low-power transformer suitable for matching an 8-ohm phone output to 500-ohm filters is the Mouser Electronics No. 42TU400, (8/500 ohm c-t) transformer with PC leads ($1.67 each).

*See Niewiadomski, part 1, September, 1985, page 17.

mismatched sources

Though transformer coupling is the easiest method of incorporating an audo filter into an existing design, it is also the most likely to result in filter mismatches. One source of mismatch is that the headphone output of a receiver may come directly from the audio power amplifier, which will have a very low output impedance — typically less than 1 ohm. The filter input is therefore most likely to be driven by a too low source impedance. This can be alleviated to a certain extent by inserting a low value resistor in series with the connection to the low-impedance winding of T1. A loss in signal will result, but it can be compensated for by

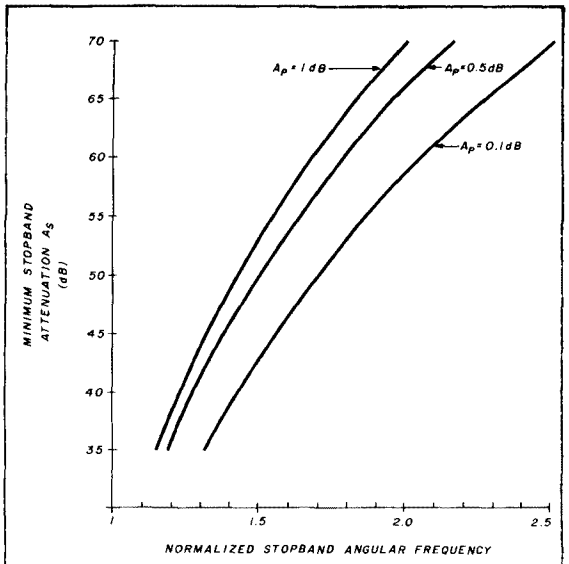


fig. 14. Minimum stopband attenuation versus normalized stopband angular frequency of 5-branch elliptic filters for various values of passband ripple amplitude (A_p).

increasing the volume level of the receiver. Another source of mismatch is the speaker or headphones themselves, which can often have an impedance which varies with frequency, thereby mismatching the output of the filter. Additional losses and mismatches can occur in the transformers. Despite these problems, a filter connected into the output of the average receiver will result in a considerable reduction in hiss and hum as well as unwanted shifted frequency SSB signals, thereby providing a corresponding increase in intelligibility.

significance of passband ripple

One notable effect on the performance of a filter using rounded component values and low-Q inductors is an increase in passband ripple. The question that has to be asked, therefore, is whether this is a significant effect in practice for a filter used in a radio receiver? I will try to answer this question by looking for other sources of audio ripple in a receiver and assessing what amplitude this ripple might typically be.

The first element which may introduce any significant ripple is the IF SSB filter. Although these filters are usually of very high quality (and expensive), they typically have a passband ripple of 1-3 dB. This ripple value may effectively be doubled if you are communicating with someone using the same model of transceiver as yours, since the peaks in ripple produced in the other station's transmit IF filter will tend to be at the same frequencies as in your receive IF filter, and vice versa.

Probably the biggest source of ripple in a receiver is the loudspeaker, which inherently has a nonlinear electrical power input to sound pressure output characteristic. Typically, a 4-inch loudspeaker in free air exhibits a 9-dB ripple from 1 kHz to 3 kHz in its sound pressure output.[4] When mounted in a metal case, with many cavities formed by metal dividers and printed circuit boards behind it, even more ripple may be produced.

Another source of ripple is the response of the human ear. Experiments performed to measure the thresholds of audibility for pure tones[5] indicate that the ear exhibits approximately 8 dB variation in the range 1 kHz to 3 kHz, being more sensitive at high frequencies.

A similar effect to that caused by the loudspeaker response will also be caused by the microphone at the transmitting end. I do not have any typical figures for this ripple, but it could be similar to the loudspeaker value.

These factors tend to negate attempts to minimize the passband ripple of an audio filter in a receiver. In this context, I believe that the passband ripple found in the filters described here is insignificant for most practical applications. Since in many cases increased passband ripple can be traded for rapid rolloff and/or greater stopband attenuation, advantage should be taken of the relative insignificance of passband ripple. This trade-off can be seen from **fig. 14**, in which curves of minimum stopband attenuation (A_S) are plotted against normalized ripple cutoff frequency for 5-branch elliptic filters of 0.1, 0.5, and 1-dB passband ripple.[3] It can be seen that for a normalized start of stopband angular frequency of 1.5 rad/sec, an extra 10 dB of stopband attenuation can be obtained by using the 1-dB rather than the 0.1-dB filter. Alternatively, the 50-dB stopband attenuation figure can be achieved at 1.42 rad/sec rather than at 1.7 rad/sec by using the 1-dB rather than the 0.1-dB filter. Another approach is to use a lower-order filter with a higher amplitude of ripple for a given application to achieve an attenuation comparable to a higher-order filter with a lower amplitude of ripple, thereby saving components and cost.

Considerable opportunity for experimentation remains on this question of passband ripple amplitude, specifically in regard to the point at which the ripple amplitude becomes both noticeable and objectionable, as well as values produced by modern receivers and transceivers.

summary

In this series of three articles, it has been shown, using simulation results, that rounding each component of an audio filter to the nearest preferred value does not result in a drastic change in performance.

Though the results obtained are not as good as with the exact component values, they would be indistinguishable for most practical purposes. This result in itself can considerably simplify the construction of filters using high-Q inductors, as exact values of capacitors and inductors do not have to be obtained or selected.

When low-Q inductors are used in filter designs, further degradation in performance takes place, but again acceptable results are obtained. Toko 10RB inductors were used in the experiments described here. Very small and inexpensive, they allow complex filters to be built compactly and inexpensively.

Several designs of elliptic low-pass filters and one Butterworth highpass filter have been described. All have been tested and shown to be thoroughly practical. Construction details have also been provided. The results obtained suggest that tables of normalized filter designs (which have for many years been used only by professionals) can now be used by Amateurs with the confidence that near-to-theoretical responses can be obtained. It has also been shown that these passive filters can be cascaded to form bandpass filters; construction details have been provided.

Attenuation equalizers have been shown to be a practical method of improving the response of filters constructed of low-Q inductors. These fairly simple networks, whose design procedure has been described, can be cascaded with the filters described here. Simulation and practical results have been presented to show the improvements obtained.

Computer simulation of filters and other networks has been shown to be a powerful design aid. As well as showing the effects of varying component values, simulation also shows that calculations such as scaling to the final cutoff frequency and impedance values can be confirmed before any components are bought.

Some methods of driving and terminating audio filters have been described, ensuring that the correct source and termination impedances for the filter are provided.

Finally, some of the factors that affect the ripple amplitude in the audio passband of a receiver were discussed. This indicated that efforts to minimize the passband ripple of an audio filter can be negated by other factors. Therefore, it can be more advantageous to trade increased passband ripple for better stopband performance at the design stage to obtain improved overall performance.

The experiments have been carried out entirely with Toko 10RB inductors, which are very compact and inexpensive. Other inductors, such as those manufactured by Dale or Miller, may also be suitable for audio filter applications.

I hope this series of articles has convinced experimenters and builders that audio filters can be constructed easily, either by using the designs presented here or by selecting other designs from tables of normalized component values. With passive designs, performance far superior to that obtainable from active designs can be obtained; perhaps passive designs will become popular again in Amateur equipment.

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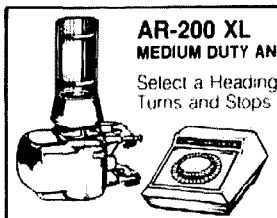
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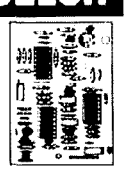
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Amateur microwave bands

My dictionary defines microwaves as the frequency range between 1 millimeter and 1 meter, the region between infrared and short-wave radio frequencies. Most Amateurs, however, consider the microwave bands those Amateur frequencies allocations above 450 MHz.

In the United States, we are presently allocated over 23,000 MHz — not counting all the frequencies above 1 mm (300 GHz). This is a huge resource. Compare this with less than 50 MHz below the microwave frequencies, of which 10 MHz are currently under attack (420-430 MHz), where greater than 99 percent of all Amateur operation takes place. For comparison, I've listed the Amateur frequency allocations above 450 MHz in the microwave bands in table **1**.

As stated in last month's column, I haven't written much about the microwave bands because I felt I had to first establish a solid base at the VHF/UHF frequencies.[1] With the past 24 columns in this series behind us, that base is now in place. With this in mind, I hope now to devote some of my columns specifically to the microwave frequencies. Therefore, this month's column will serve as sort of an introduction to the microwave bands, frequencies in use, and activities thereon.

table 1. Major worldwide Amateur microwave frequency allocations. Many countries share the same frequency bands, although the frequencies may be shifted slightly.

band	frequency	notes
33 cm	902-928 MHz	Available in USA since Sep. 28 1985. Canada has same but on A3/F3 only.
23 cm	1215-1300 MHz	1215-1240 still available in Canada, but recently withdrawn in USA. Some countries in Region 1 do not have the full assignment. Others have power or EIRP restrictions.
13 cm	2300-2450 MHz	2310-2390 was removed from USA use on November 6, 1984. Many Europeans cannot operate below 2320. Japan has only 2400-2450.
9 cm	3300-3500 MHz	Some area restrictions apply. UK and West Germany have 3400-3475.
6 cm	5650-5925 MHz	Some area restrictions apply. UK and West Germany have 5650-5850.
3 cm	10.0-10.5 GHz	
12 mm	24-24.25 GHz	24-24.05 in West Germany
6 mm	47-50 GHz	47-47.2 in West Germany, 48-50 in USA.
4 mm	71-76 GHz	75.5-76 in West Germany
2 mm	142-170 GHz	165-170 GHz in USA, 142-144 GHz in West Germany.
1.2 mm	240-250 GHz	248-250 in West Germany
1 mm	300 GHz and above	No restrictions in USA.

a look back

First, let's take a glance at the past. Amateurs have really only operated on the microwave frequencies since World War II.[2] Most microwave operation in North America has been conducted on the 23 cm (1215-1300 MHz), 13 cm (2300-2450 MHz), and 3 cm (10-10.5 GHz) bands with a little activity on 9 cm (3300-3500 MHz), 6 cm (5650-5925 MHz), 12 mm (24-25.25 GHz) and 6 mm (47-50 GHz) thrown in. In fact, the only reported North American Amateur contacts above 50 GHz are in the optical frequency bands, those frequencies above 300 GHz![3] I've updated the DX record table and the latest records (as of this date) are shown in tables 2, 3 and 4.

During World War II, the military classified frequencies by assigning letter designations to the various bands. Some of those designations are still in use today, especially in reference to older surplus gear. **Table 5** shows these designations. However, tri-Service designations have now changed to a more orderly list; the more recent band designators are shown in table **6**.

Until recently, most Amateur gear used on the microwave bands was commercial or surplus equipment that

table 2. Worldwide claimed microwave terrestrial DX records.

frequency	record holder	date of QSO	prop mode	dx miles (km)
1296 MHz	KH6HME-N6CA	6/24/84	Tropoducting	2472 (3977)
2.3 GHz	VK5QR-VK6WG	2/17/78	Tropoducting	1170 (1883)
3.3 GHz	G3LQR-SM6HYG	7/11/83	Ducting	576 (927)
5.7 GHz	G3ZEZ-SM6HYG	7/12/83	Ducting	610 (981)
10 GHz	I0SNY/EA9-I0YLI/IE9	7/08/83	Ducting	1032 (1660)
24 GHz	I3SOY/3, IW3EHQ/3-I4BER/6, I4CHY/6	4/25/84	LOS	180 (289)
47 GHz	HB9AMH/P-HB9MIN/P	6/11/84	LOS	33 (53)
474 GHz	K6MEP-WA6EJO	6/09/79	LOS	15 (24)

table 3. North American claimed microwave DX records by propagation modes. The records are listed alphabetically by mode. Ducting is suspected where the paths are mostly over water. No efforts are made to separate ducting on overland paths, so they're grouped under "tropo".

frequency	record holder	date	prop mode	dx miles (km)
1296 MHz	KH6HME-N6CA	6/24/84	Ducting	2472 (3977)
	K2UYH-VK5MC	12/06/81	EME	10,562 (16995)
	W4WSR-WA5TKU	6/03/85	Tropo	1112 (1790)
2.3 GHz	PA0SSB-W6YFK	4/5/81	EME	5491 (8836)
	W4HHK-W8YIO	7/28/83	Tropo	583 (938)
3.3 GHz	K6HIJ/6-W6IFE/6	6/18/70	LOS	214 (344)
5.6 GHz	K5FUD-K5PJR	9/20/77	Tropo	267 (430)
10 GHz	WA4GHK/4-WD4NGG	8/07/84	Ducting	297 (478)
	W7JIP/7-W7LHL/7	7/31/60	LOS	265 (426)
24 GHz	KX0O/0, W0MXY/0-NK0P/0, WA0VSL/0	8/24/85	LOS	74 (119)
48 GHz	W2SZ/1-WA2AAU/1	9/08/84	LOS	0.3 (.5)
71-250 GHz	None reported			
Above 300 GHz	K6MEP-WA6EJO	6/09/79	LOS	15 (24)

table 4. Worldwide claimed EME Microwave DX records.

frequency	record holder	date of QSO	prop mode	dx miles (km)
1296 MHz	PA0SSB-ZL3AAD	6/13/83	EME	11595 (18657)
2304 MHz	PA0SSB-W6YFK	4/05/85	EME	5491 (8836)
3300 MHz and above: None reported.				

was often borrowed or modified by Amateurs for their use. The primary modes of communication were often wideband FM and pulse.[2]

The San Bernadino Microwave Society (SBMS) was an early supporter of the microwave bands with their "polaplexers."[4] This scheme uses only one oscillator at each end of the path that is modulated for transmitting and provides the LO for receiving. This LO was often a klystron; in more recent designs, it's often a Gunn diode oscillator.

Polaplexers were primarily used to set DX records between mountain tops using portable gear — with "portable" meaning gear that could be carried by two men and a boy! Indeed, the SBMS 9 cm record for North America still stands (**table 3**). The polaplexer's days seem to be numbered, and the gunnplexer is rapidly replacing it on the 3 cm and 12 mm bands.[5]

looking forward

Before 1970, many of the microwave records were held by North Americans. Then the British and Germans started to take on the challenge. Before long, many of the European countries were there in great numbers. When you look at the worldwide DX records in **table 2** and compare these with the North American records in **table 3**, you must conclude that there's plenty of room for challenging DX records. Furthermore, I'm not aware of any successful Amateur EME activity above the 13 cm band (**table 4**).

At the present time, narrow-band modes such as CW and SSB are starting to take hold even on 3 cm, a hotbed of activity in the UK. The state of the art advances rapidly. Needless to say, the Europeans have a good track record. So let's get in on the ground floor and see what we in North America can do!

Unless you're a radio astronomer, it's no fun to listen to white noise. Activity on the microwave bands, especially above 13 cm, will be successful only if individuals band together and concentrate their efforts. We have the technology — all we need to do is spend the time and make the effort.

The FCC has finally allocated the 33 cm or 902-928 MHz band to radio Amateurs in the USA on a secondary basis. Canadian Amateurs already

have the band, but only for A3J emission. What a great challenge this band offers! Gear for this band is easier to design and more readily available than gear for the upper microwave bands.

microwave frequencies and modes

Most of the North American activity below 9 cm is now on CW and SSB. Many years ago weak signal operators used multipliers from 144 MHz to go to the 70 cm and higher bands. As a result, the weak signal frequencies of 432.0 MHz (144 X 3), 1296 MHz (144 X 9), 2304 MHz (144 X 16), 3456 MHz (144 X 24), 5760 MHz (144 X 40), and 10368 MHz (144 X 72) became established and are still recognized worldwide.

Because CW and especially SSB will probably become widely used, the transmitters and receivers presently employed must be upgraded. Some Gunnplexer operation on 3 cm and 12 mm will undoubtedly continue. But these units will probably be upgraded with frequency/phase locking, improving frequency stability and hence weak signal operation.

The prime weak signal calling frequencies on the lower microwave bands have mostly shifted to 1296.1 and 2304.1 MHz to allow EME exclusively between 1296.0-1296.05 and 2304.0-2304.05 MHz. EME on the 13 cm band will be somewhat fragmented since many European Amateurs can no longer operate below 2320 MHz, and the FCC has removed 2310-2390 MHz in the USA. Therefore, cross-band operation (2304.0 to 2320.15 MHz) is now in use. The ARRL VUAC has recommended 903.1 MHz as the weak signal calling frequency on 33 cm, with EME between 903.0 and 903.05 MHz.

propagation

Radio propagation on the microwave bands is somewhat different than on the VHF/UHF bands. Aurora, while theoretically possible, has yet to be reported. Meteor scatter, sporadic E, F2, and FAI are theoretically impossible on these frequencies.

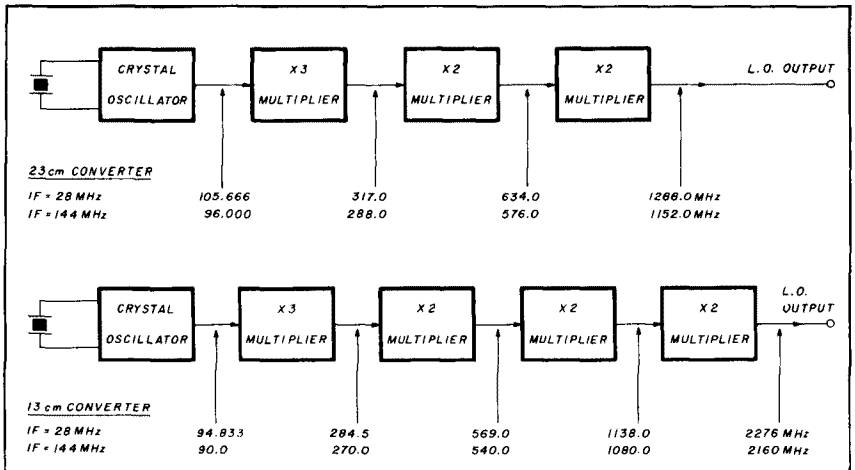


fig. 1. Recommended multiplier scheme for a 1296 and 2304 MHz converter/transverter. (See text for further explanation.)

The most common modes of microwave propagation will probably be line-of-sight (LOS), tropospheric bending, tropospheric ducting, scatter, and EME. Scatter will undoubtedly become more popular for DX since it's very prevalent at microwave frequencies. Large stationary objects (buildings, hills, mountains, and such) as well as aircraft and lightning make excellent scattering media.[3] Scatter is often overlooked, but even local operation may be enhanced between stations that have obstructed paths if each station aims at some large hill, building, TV tower, or other structure that they can both "see."

The biggest impediments to DX on microwaves will be foliage/evergreen attenuation, feedline losses, low transmitter power, and narrow antenna beamwidths. The latter problems can all be improved with continued development in state-of-the-art equipment (see below).

Therefore, with the exception of EME, the DX records on the microwave bands will probably be shorter than the VHF/UHF records until better equipment becomes available and activity increases. For more information on the subject of propagation, refer to references 3 and 6.

receiving equipment

Almost all microwave weak signal operation now uses crystal controlled converters, with 28 and 144 MHz IFs most popular. This permits most of the weak signal types of emissions to be used.

Adequate RF filtering, especially at the image frequency, is required for low-noise operation. Allowing other RF frequencies outside the band of interest to enter the receiver lowers performance, especially if the signals are strong. If a 28 MHz IF is used, filtering is more critical.

However, simple filter designs such as the interdigital type are now available to Amateurs.[7,8] These filters are particularly easy to design using computer programs that can change the mechanical and electrical specifications at will. Construction is not beyond the hand tool level.

The present state-of-the-art permits noise figures below 3 dB up to 25 GHz. Only a few years ago this would have been unheard of except with the use of parametric amplifiers and masers, both out of the reach of most Radio Amateurs.

Microwave bipolar transistors are now inexpensive and can deliver 1 to 3 dB noise figures through 4 GHz. Affordable (less than $25) GaAs FETs can now provide less than 2 dB noise figures through 3 cm.[9] The lower noise figure high mobility electron transistors (HMETs) are coming.[10] Although they

presently cost around $150.00 (if they can be obtained at all), they reportedly will eventually be less expensive than GaAsFETs since they're easier to manufacture.

The local oscillator (LO) of a microwave converter is very important. As I've stated in previous columns, the basic oscillator for VHF/UHF and above should be an overtone crystal type in the 90-120 MHz region.[11,12] Not only will this reduce phase noise, but it will improve stability, generate fewer spurs, and require less multiplication.[13] As a result, there will be fewer birdies and fewer LO products to be filtered out.

Multipliers should also be clean. I recommend that doublers be used if possible since they're more easily filtered and more efficient. If triplers are used, they should be limited to about 400 MHz on the output. Some recommended combinations for 1296 and 2304 MHz are shown in **fig. 1.** Many types of mixers have been used over the years. The through-line single-diode type was commonly used on microwave frequencies in the early days.[2] More recently, interdigital mixers and the anti-parallel harmonically pumped mixers have been used, since they require a lower frequency LO at a sub-harmonic of the desired frequency.[14,15,16]

However, I prefer the balanced mixer types.[11,12] They're usually broadband and represent 50 ohms at all three ports (LO, RF, and IF). Hence they're easy to work with, especially in a modular converter. The double-balanced (DBM) type with four diodes is preferred, and are becoming quite easy to obtain and are reasonable priced (usually $10-40 for 1-4.5 GHz), with the MDS and TVRO market paving the way for the 2-5 GHz spectrum. The DBM usually requires about 5 milliwatts of LO power.

Finally, the post amplifier (the one following the mixer) should have a low (1 to 2 dB) noise figure to lessen the preamplifier gain requirements. A suitably low noise figure bipolar preamplifier that will cover any IF from 14-150 MHz is described in references 12 and 17. Other low noise circuits are also acceptable, especially those with good input VSWR.

transmitters

Multipliers are fine for the microwave bands. As mentioned before, 144 MHz transmitters used to be multiplied all the way up to the desired microwave band. However, their frequency stability left something to be desired.

If you decide to use a multiplier scheme, there's a preferred method that has been used by many microwave and contest stations.[18,19] Basically, it uses a 96 MHz oscillator which is multiplied up to 1152 MHz. This is a convenient frequency since its second, third, fifth, and ninth harmonics allow operation on 2304, 3456, 5760, and 10368 MHz respectively, as shown in **fig. 2.** Coincidentally, 1152 mixed with 144 MHz yields 1296 MHz as a bonus!

However, the transverter is rapidly becoming the most popular scheme for getting on the microwave bands. Since only one LO is required per band, it significantly lowers the cost and reduces complexity, not to mention construction time. Properly built, it will have excellent frequency stability. In addition, a transverter allows CW as well as SSB operation at the flip of a switch on the exciter.

The only problem with a transverter is that a separate oscillator is required for each microwave band, unlike the 1152 MHz transmitter scheme just mentioned. Many simple two- or three-band schemes have been proposed to lower the number of required oscillators.

Recently a more or less universal microwave scheme was proposed using a single oscillator.[20] Although it may at first seem complicated, it would make a dandy add-on type of building block, since once the first block is in place, bands can be readily added. The disadvantage of the scheme is that a separate converter is required at 288 MHz. However, that frequency is a snap to work with and will be high enough to make image rejection a trivial problem. The basic scheme is shown in **fig. 3.**

table 5. Major microwave band destinations used during and after World War II. Note that there are some overlaps to add to the confusion.

frequency (GHz)	designation
0.10 - 0.225	G
0.225 - 0.390	P
0.390 - 1.55	L
1.55 - 5.2	S
3.9 - 6.2	C
5.2 - 10.9	X
10.9 - 36	K
12.9 - 18	Ku
26.5 - 40	Ka
36 - 46	Q
46 - 56	V

table 6. Modern Tri-Service microwave band designations in present use.

frequency (GHz)	designation
0 - 0.25	A
0.25 - 0.5	B
0.5 - 1	C
1 - 2	D
2 - 3	E
3 - 4	F
4 - 6	G
6 - 8	H
8 - 10	I
10 - 20	J
20 - 40	K
40 - 60	L
60 - 100	M

obtaining power

Power amplifiers aren't always required on the microwave bands. If solid-state multipliers are used, up to several watts can be generated on the lower microwave bands.

Solid-state bipolar transistors can generate several watts of power up through 6 GHz. Power MOSFETs that generate several watts at 1 GHz have been demonstrated. Power GaAs FETS that can deliver 5 to 10 watts of output through 10 GHz are also available.

Until recently, the 2C39/7289 has been the king of transmitting tubes on the lower microwave bands. Single, two, four, six and even eight-tube

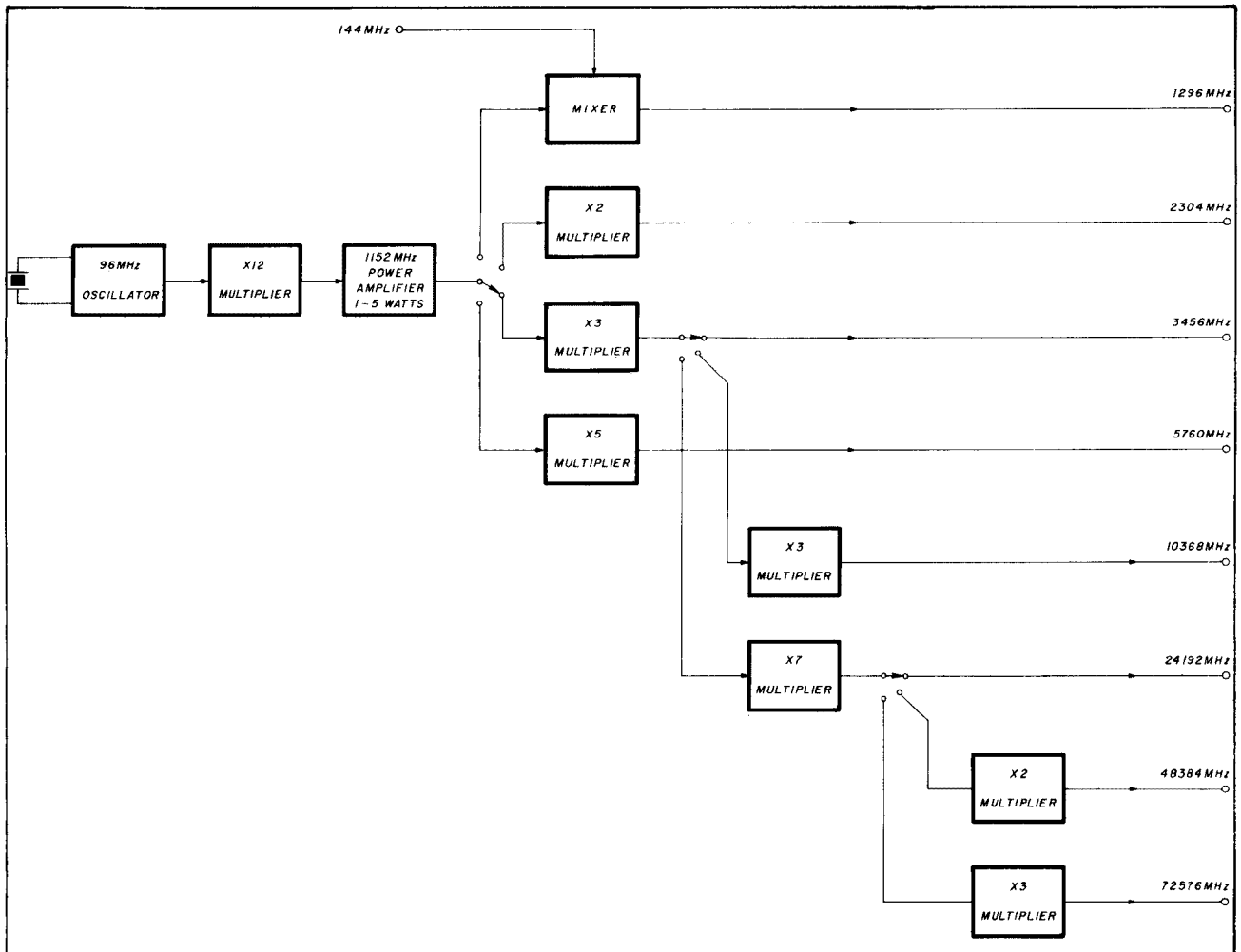


fig. 2. A frequency multiplication scheme for the Amateur microwave bands. (ref. 18 & 19).

amplifiers have been used through 2304 MHz.[21-25] The eight-tube amplifier design described in reference 24, however, is not recommended, because this design offers very low efficiency (30-40 percent).

Recently a new single-tube 23-cm amplifier design using a UHF TV tube was introduced. It will reportedly deliver over 400 watts with a 10 dB gain.[26] Other tubes and designs are available. Traveling wave tubes (TWT) will deliver up to 100 watts at microwave frequencies, although those above 10 watts are rare in Amateur circles. The outstanding properties of the TWT are that it is broadband (typically one octave), has good linearity, and 25 to 35 dB of gain is typical!

Amateurs are also using high-power klystrons, but they're difficult to obtain and quite high in price. Like TWTs, they frequently have gains of over 30 dB. Magnetrons are another possibility. If someone can find out how to use the type presently so common in microwave ovens, we could have a readily available, low-cost high power amplifier source for 13 cm!

antennas

The principal types of antennas used on the microwave bands are the Yagi, loop Yagi, horn, and the parabolic dish. Because of the critical dimensional requirements, the Yagi will probably not be widely used except below 1300 MHz. But the loop Yagi, with its unique mounting and matching methods, should be a good performer where 19 to 21 dBi of gain is required with low wind resistance up through at least 3500 MHz.[27, 28] The microwave horn has only moderate gain (10 to 20 dBi), but is widely used on 3-cm gunnplexers and as a feed system for parabolic dishes where low gain or controlled beamwidth is required. The parabolic dish, with its high gain, will probably remain the standard antenna for the microwave bands because it's easily designed, has predictable gain, and is easy to feed.[29]

feedlines

The microwave bands are a transitional region for feedlines.[30] Coaxial

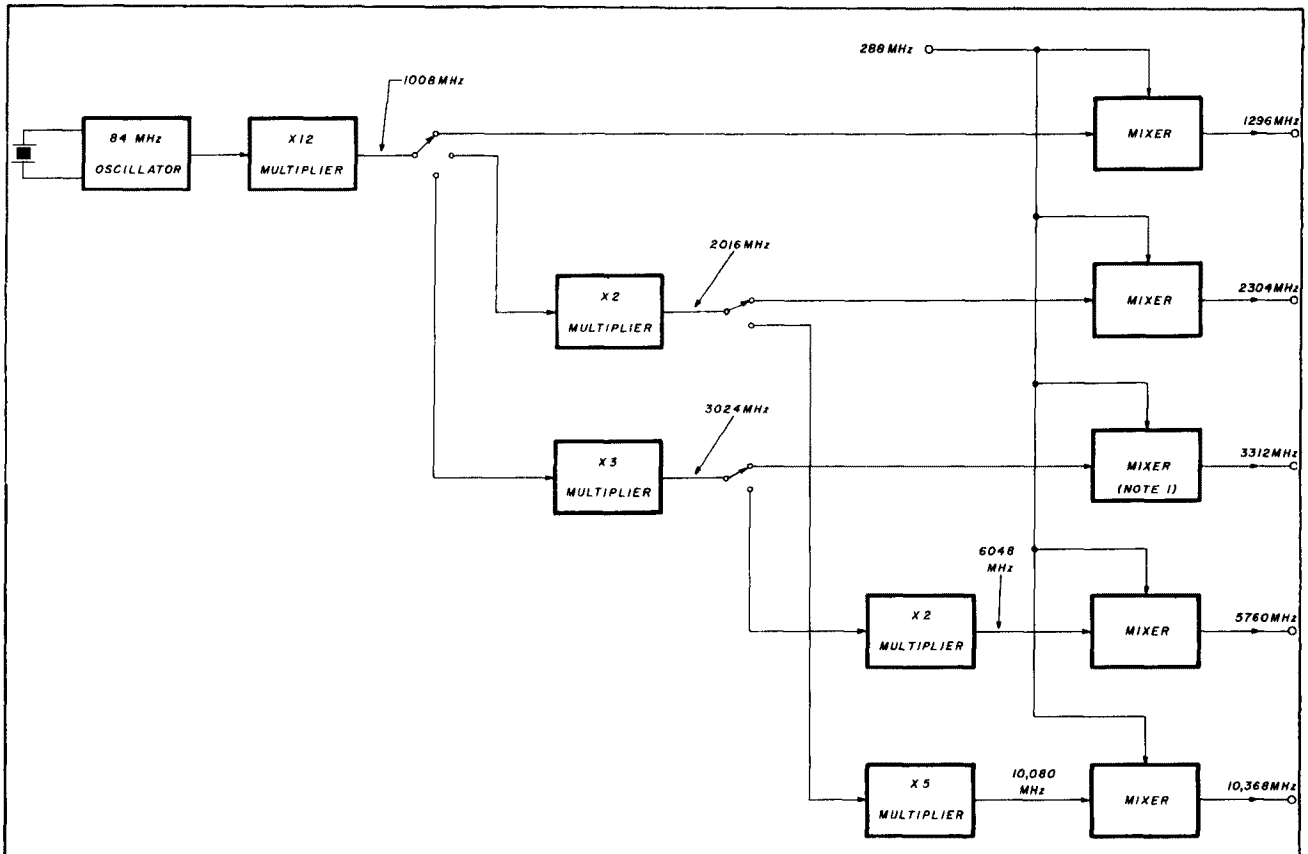


fig. 3. A universal frequency conversion scheme for many of the lower Amateur microwave bands. (ref. 20). Note 1: If 432 MHz is injected instead of 288 MHz, the output will be on 3456 MHz, the standard weak signal frequency.

cable is improving all the time; it's easy to work with, especially where bends or flexibility are required. It will probably remain the most popular line at the lower microwave frequencies and slowly move up in frequency as the state-of-the-art improves. In this regard, the air dielectric types of coax such as Heliax™ are preferred.

Waveguide will probably remain the most common type of transmission line on the microwave bands above 10 GHz. It's readily available as surplus and has very low insertion loss. Don't overlook the possibilities of "G-line," which has the potential for low loss at a very moderate cost — and you can "roll your own."[30]

commercial gear

Those who know me well are aware that I prefer to build my own gear from scratch because it allows me flexibility in choosing features as well as the ability to make modifications and improvements without the fear of lowering resale value. It also yields a great sense of accomplishment when everything works!

However, there's now a fair amount of microwave gear available not only from Amateur suppliers, but from commercial suppliers, too. Furthermore, because the state-of-the-art in commercial equipment is moving ahead so rapidly, there's a lot of surplus equipment available to the Amateur. Check the Amateur magazines for further information. The microwave bands are no longer the exclusive territory of the homebrewer!

reference material

There's plenty of material about the microwave bands available if you know where to look.[31] Of particular note are the RSGB's *VHF/UHF Manual*, *The Microwave Newsletter Technical Collection*, *The Gunnplexer Cookbook*, *VHF Communications*, *DUBUS*, *IEEE Microwave Theory Transactions*, and *The UHF Compendium*, for example. (For information on how to obtain any of these, see reference 31.) Bibliographies are also available.[32, 33, 34]

Finally, numerous trade magazines are available for those engaged in the electronics field.[31] The material is available, but you'll have to take the initiative!

summary

In this month's column, I've tried to provide an introduction to the microwave bands. Because complete examination of the topic would obviously be beyond the scope of this column, I strongly advise you to refer to the references cited.

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The Amateur microwave bands are an exciting area for exploration and challenge. Surely they'll be the resource of the future — not just for casual chit-chat, but as *the* place for improving the state-of-the-art in communications and equipment design.

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important VHF/UHF events:

January 3:	*Peak of Quadrantrids Meteor Shower predicted at 2300 UTC*
January 8:	*EME perigee*
January 11-12:	*ARRL VHF Sweepstakes Contest*
February 4:	*EME perigee*

ham radio

short circuit

VHF/UHF world

In the September, 1985 column, "VHF/UHF World" (page 45), the 2 7/8-inch dimension shown in **fig. 2C** should be changed to ±2 1/2 inches. This is the height of the loop (from top to bottom), not the diameter, and will cause the loop to appear somewhat oval rather than round.

During final adjustment, the loop Yagi should be tested for minimum VSWR. If it is not below 1.2:1 (after adjusting the spacings between the first director and the reflector, as explained in the text) adjust the loop height carefully to obtain the best match. After adjustment, the junction point where the driven element connects to the brass bolt and the transmission line should be thoroughly soldered together.

PRACTICALLY SPEAKING ... Joe Carr K4IPV

troubleshooting I: signal tracing/injection

SCENE: It's one of those evenings when you can't determine whether the band has just died a little earlier in the evening than usual or somebody's stuck a pin through your antenna coax. Wise Old Elmer (WOE) is leisurely tuning across 15 meters when the doorbell rings. He opens the door. Standing there, looking more than a little crestfallen, is Young Squirt (YS), a novice who has only recently upgraded to General Class.

"Hiya, Young Squirt — what's the matter, a truck run into your tower?" Young Squirt replies, "Almost. My $800 Super Band-Buster Single Sideband (SBBSSB) transceiver died on receive — think ya' can help?"

Now, WOE is one of those old-fashioned people who believes that if you give someone a fish then they eat for today, but if you teach them how to fish, then you feed them for a lifetime. "Well, Squirt, I reckon I could fix it — after all, you bought the shop manual and I got a workshop full of test equipment. But what good does it do if I fix it?"

WOE pauses, knowing YS's objection because he remembers a scene like that, oh, so many years ago, back when he thought the QCWA met in the local nursing home.

"Well, answers YS, ...it'd get me back on the air."

"Young Squirt," says WOE, assuming a fatherly tone, "you just passed a community college course in radio and nearly aced the General class

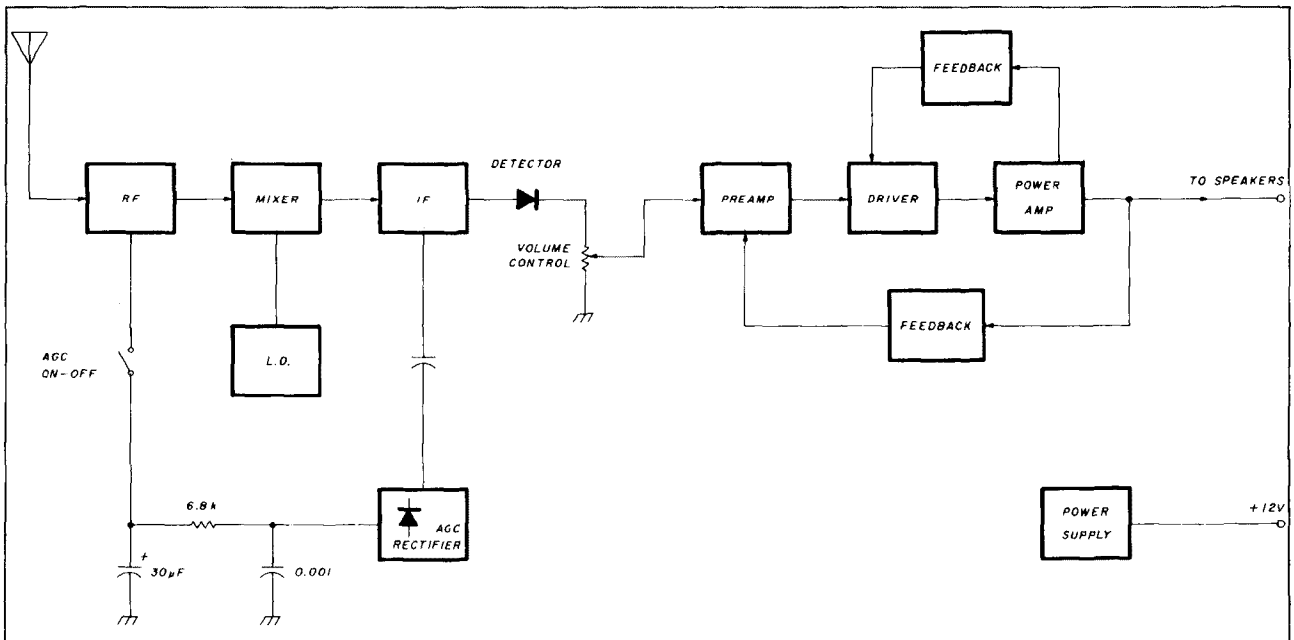


fig. 1. Standard radio receiver block diagram.

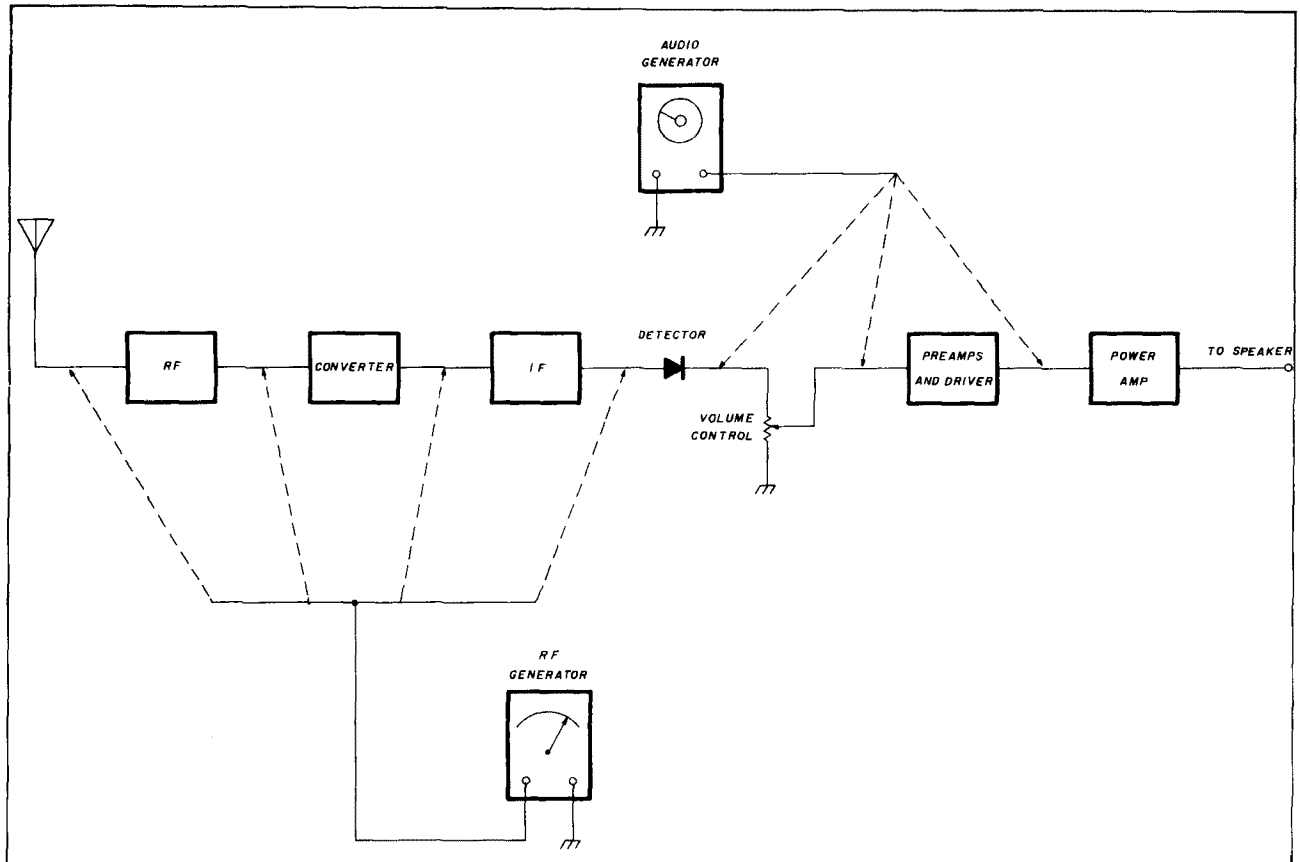


fig. 2. Circuit of *fig. 1* (standard receiver) redrawn to show signal injection.

license exam. Why don't I show you how to troubleshoot that rig of yours, and see if you can learn a thing or two?"

YS is startled at the suggestion. Elated, but still questioning his own ability, he admits that WOE knows a lot more than he does, so puts his anxiety on hold. But when WOE heads over to the workbench — muttering something about getting out his fishing pole — the anxiety returns.

With the transceiver on the workbench, WOE directs YS to remove the top and bottom covers according to instructions in the service manual. Taking out a piece of paper and a stubby No. 2 pencil, he draws a block diagram of a standard radio receiver (**fig. 1**).

"Take a look at this picture, Squirt. Note the various stages. We have an RF amplifier tuned to the incoming signal, a converter or mixer/local-oscillator stage that converts the RF signal to the IF frequency, an IF amplifier where most of the gain and selectivity are placed, a detector — let me draw a simple AM envelope detector, just for the sake of argument — and an audio amplifier chain. There might be various feedback loops in the audio, but we can leave them out for now. There might also be an automatic gain control (AGC) circuit. This circuit is simple compared with the SBBSSB receiver section, but it serves to show you how to bait your hook."

"Huh? Bait my hook?"

WOE smirks. "Let's talk a little about troubleshooting philosophy, YS," he says. "It'll help us think this process through. *The vast majority of electronic troubleshooting problems involve either losing a desired path for current or gaining an undesired path for current.*" That certainly sounds simple, so Squirt figures WOE just might know a thing or two. "Furthermore," WOE continues, "there is, in each stage, an AC path for current and a DC path for current. Although there are cases in which examining the DC path in each stage will allow us to isolate the defective section of the receiver, it's more often the case that the signal path — an AC path — is the best way to attack the problem."

"DC path, AC path?" YS asks, "that sounds familiar. I talked to Sally Solderburn last week, and she told me about using those ideas to troubleshoot solid-state circuits." (Sally only recently obtained an Amateur license, but had been the Chief Engineer of the local National Public Radio FM station in their town for several years. She'd also spent several years repairing CB and taxicab two-way radios.)

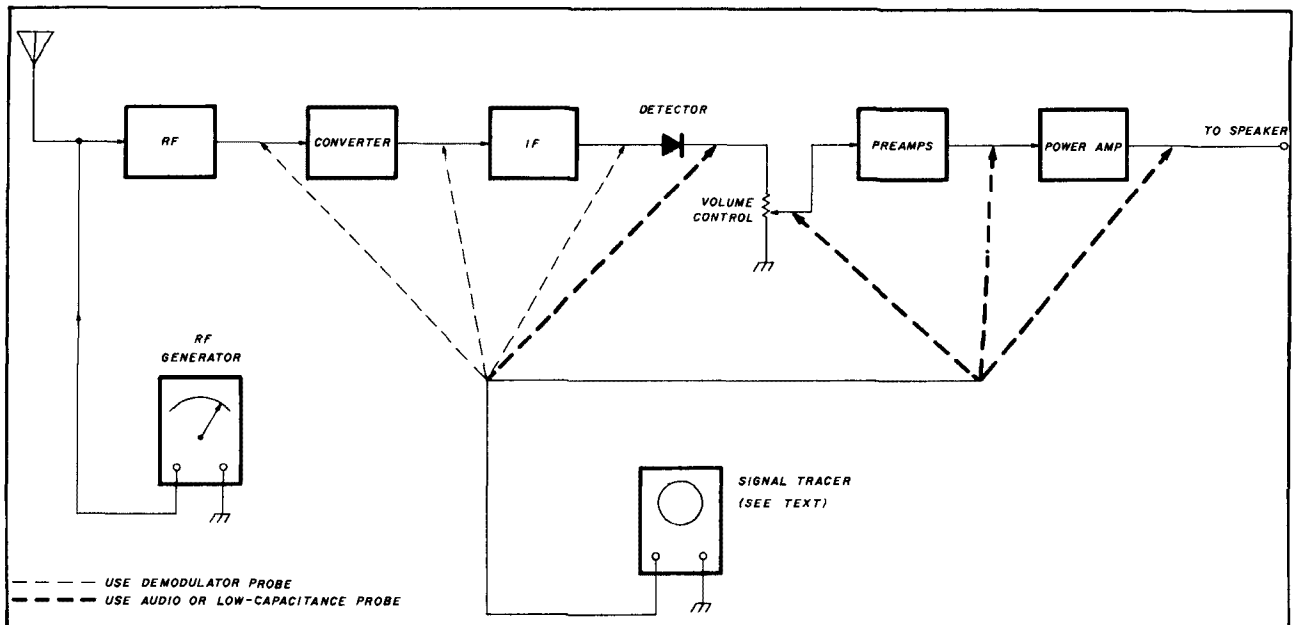


fig. 3. Circuit of *fig. 1* (standard receiver) redrawn to illustrate signal tracing.

"Yeah, Sally showed me a thing or two about troubleshooting solid-state circuits," replies WOE, "but let's get to that in a moment. The first job is to isolate the dead stage. There are two methods we can use: signal tracing or signal injection."

Taking out another sheet of paper, WOE redraws the circuit of fig. **1** in the form of fig. **2**. "In signal injection we use a signal generator to insert a signal into the various stages of the receiver, while monitoring the output signal on the speaker or a pair of headphones.

"In the textbook case — I use that term because, practically speaking, there are telling indications that allow us to cut the process short — we'll first use an audio generator at the input of the audio output stage. If this signal creates an appropriate response in the output — that is, a tone in the speaker — then we move the signal generator back to the input of the AF preamps or the volume control. If the audio stages prove out OK, then we have to switch to an RF signal generator. At first, set the RF sig gen to the IF frequency. Inject the signal into the input of each IF stage (there are three stages, you know) and the output of the converter (or mixer, if that system is used). Finally, if all these stages prove out, as indicated by the signal getting all the way through to the output, switch the sig gen frequency to the RF frequency. This step's a little tricky, in some cases, because the receiver and sig gen aren't tuned to the same frequency; the dials are way off from each other. Rock either the receiver or signal generator dial back and forth to find the signal. The last stage is the RF amplifier."

YS marvels at the simple logic of stage isolation, but now wonders what "signal tracing" is: "Tell me, Wise Old Elmer," he asks, "you mentioned signal tracing a couple minutes ago — what's that?" WOE reaches for his old Heathkit signal tracer and shows it to YS. "Well, Squirt," he begins, "the signal tracer is merely a high gain audio amplifier with a loudspeaker output. The input is a switchable probe. In one switch position the probe acts like a simple low-capacitance probe, while in the other it's a demodulator probe. This latter position will demodulate an AM signal." Turning the now well-used piece of paper over, WOE again draws the simplified block diagram of a radio receiver (fig.**3**) and shows Squirt how to signal trace.

"Remember how in signal injection we started at the audio output stage and worked backward toward the RF amplifier?" he asks. "In signal tracing, we work in the *opposite* direction — from the RF amplifier toward the AF stages. It's important to have the receiver tuned to some reasonably strong signal that will stay on the air for a little while. On some Amateur bands there are enough signals so that this is not a problem — 75 meters after 6 PM for example — but in other cases, we'll have to supply a signal. In our area, WWV on 10 MHz is usually strong, so we can use the WWV mode of the receiver. Otherwise, simply connect a signal generator to the antenna terminals. Set the output frequency to the same frequency as the receiver dial.

"Use the signal tracer in the 'demodulator' mode (also called the 'RF' mode on some instruments). Look for the signal first at the output of the RF stage, then at the converter (or mixer), at each IF stage in succession, and finally at the input of the

detector. If the bad stage is not found, then switch to the AF mode on the signal tracer, and continue to the detector and each of the audio stages in succession."

"The process of stage isolation," WOE continues, "is used to limit the problem area to a single bad stage. Once we determine which stage is dead, then we can use other methods to figure out which component is defective."

YS is impressed. "Darn!" he exclaims, "that's simple, WOE. Why didn't I think of that?"

"Hmmmm..." thinks WOE, "that's a good question."

"WOE," asks YS, "you mentioned some shortcuts...what were you talking about?"

"Well, I don't mean to imply that the process isn't valid, because it works in almost all cases — in fact, it's never failed for me. But a little experience shows us that some indications allow us to eliminate certain stages. You know, professional technicians often make more money from merely observing sets than from testing them. People like your friend Sally make their living by fixing as many sets per working hour as possible, so they're deeper in the bucks if an educated guess shortens the process."

Squirt is now almost breathless — but whether from a desire to impress Sally or from the technical wisdom he's gaining isn't clear. "What kind of shortcuts are there?" he asks.

"Well, let's take a quick look at your rig and see." WOE makes sure the rig is turned off and in the receive mode. He plugs in the AC power cord and turns on the power switch. "Run the AF gain control up and down through its range several times, YS," he directs. "Hear that scratching? Very few volume controls are perfectly clean potentiometers. Almost all of 'em have enough scratching to let you know whether or not the AF stages are working."

YS turns the control several more times, and sure enough, the speaker emits a rough scratching sound as the control is adjusted.

"That scratching tells us that the AF amplifiers are probably OK," WOE explains. "Now set the AF gain at or near maximum and turn the RF gain control all the way up. Notice that the set is all but dead — there's only a small hissing sound. That tells us that the problem is probably in the detector, IF, or mixer/oscillator (or converter) stages. If the mixer/oscillator and all following stages are working, and the problem is in the RF amplifier, then we'd hear a loud hiss amplified by the high gain IF amplifier. This hiss is a combination of thermal noise in the stages plus a certain amount of noise created by the heterodyning process. Listening to the receiver with the antenna disconnected is nearly the same sound."

WOE takes out another sheet of paper, brushes away some instant coffee crystals, and makes a list for his young friend. He titles it "common faults and their indications, especially for AM/SSB radios" and begins to write

1. With all gain controls turned up, a loud hiss indicates a lot of gain is present, so the problem is closer to the front-end.

2. Picking up only a few stations in a normally crowded band (not 15 meters at 10 PM) could indicate a sensitivity problem. If there's not a lot of hiss, then look in the IF amplifiers (especially the early stages). The fact that the signals are there, in the right place, exonerates the mixer/oscillator.

3. A lot of signals, but with low volume, usually indicates that one of the latter IF stages, the detector, or one of the AF stages is faulty. If the volume control scratch is normal, then the AF stages are probably OK.

4. Tune the radio across a busy band while watching the S-meter. If there are periodic deflections — indicating that the receiver tuned across a signal — but there's still no output signal, then the dead stage is between the S-metered stage and the output.

5. A constant — or nearly constant — indication on the S-meter might indicate an oscillating IF amplifier or other stage.

"Well, there you have it, Squirt. Think you can show me which stage is bad?"

YS, who by now has had both his consciousness and confidence raised, mutters, "I expect so."

next month

This month we used a little "Elmer Scenario" to illustrate a little lesson in troubleshooting. Next month we'll return to our usual format and tell you what Sally Solderburn taught YS about troubleshooting.

have a question for Joe Carr?

Send your question to Joe Carr, *ham radio*, Greenville, New Hampshire 03048. While not every letter can be answered personally, he will try to answer as many as possible in this column. —Ed.



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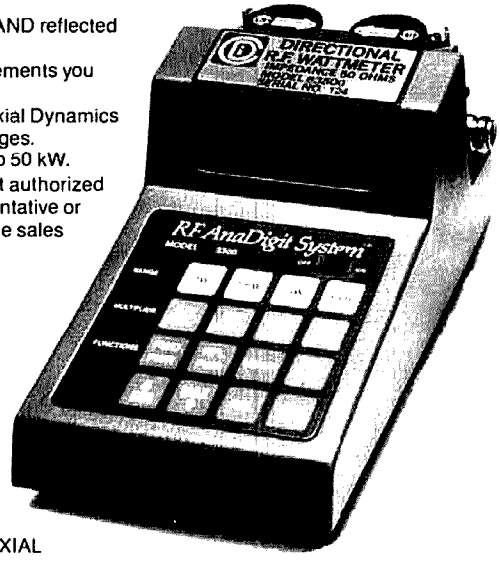
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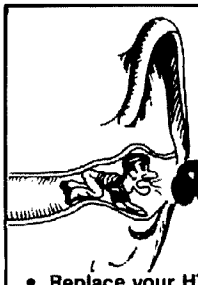
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short circuit

K1BQT transceiver

The following corrections should be made to K1BQT's November article, "A Compact 75-meter Monoband Transceiver":

AUDIO AMPLIFIER/AGC MODULE: Schematic omitted 1N914 in line from audio output to gate of Q1. Also, AGC delay may be too long with 2.5 μF on gate of Q1. If so, substitute 1 or 1.5 μF.

DRIVER/BPF MODULE: Q3 gate No. 1, resistor value should be 10k. Schematic omitted 0.01 between L4 and gate No. 1. L4 is direct coupled — *not link coupled* — to Q3. Also, schematic omitted RFC1 on Q4.

EXCITER MODULE: Schematic omitted 0.1 bypass on U5 pin 3. Delete 1k resistor shown between mic gain and U6 pin 1.

RECEIVER MODULE: Schematic omitted 0.1 bypass capacitors on U1 pin 7, +R, and U3 pin 7. On PCB, move 0.1 bypass from ALC input to U2 pin 5.

Finally, use silver mica capacitors in the transmitter BPF. Parasitic oscillation of the driver was traced to inferior quality ceramic capacitors in the two units.

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ham radio TECHNIQUES

Bill Orr W6SAI

let's hear it for the good guys

I have noticed a slight tendency, on the part of some high-tech people, to look down on hams and ham activity — as if such a hobby somehow lessened the professionalism of those whose livelihood is RF engineering. I challenge that notion. Quite to the contrary, experience as a practical operator cannot but deepen one's understanding of electromagnetic wave generation and propagation. In fact, if there are any computer kids among our readers who are not hams, we urge you to give it a try. It will give you much of the education, in this strange "new" field, that you're looking for.

These interesting and encouraging words appeared in an editorial by Keith Aldrich, publisher of *rf design* (August, 1985).*

Other Radio Amateurs in the technical fields have noticed the professional "snub" on occasions, but this is far outweighed by the great number of professionals that *are* hams. All one has to do is attend the Radio Amateur dinner and reception at the National Association of Broadcasters Convention, or the Amateur luncheon held at the annual gathering of the Armed Forces Communications and Electronics Association. You'll find an eye-popping number of top-notch professionals — who are also enthusiastic Radio Amateurs — at these meetings! So, let's hear it for the good guys and for Keith Aldrich, who I hope will become a ham himself one of these days!

the phasing SSB system — born again!

Old-timers will recall the phasing system of SSB. Remember the Central Electronics 10A exciter? This thought should bring tears to the eyes of any sidebander who was active in the early 1950s. The phasing system died a quick death with the advent of the latest edition of the *ARRL Handbook*. (A friend of mine chided me for including a short section on the phasing SSB system in the latest edition of *Radio Handbook*. My reply to him was that it is still a valid system of generating a sideband signal and possesses some virtues that the filter system does not have. His only reply to me was a look of pity.)

But now it looks as if the phasing system has finally come into its own, especially in the field of VHF/UHF operation.

A major disadvantage of SSB is that it requires relatively inefficient linear amplifiers to reach high power levels. Unfortunately, reasonable efficiencies at VHF and higher are obtainable only with class C amplification, which precludes the use of a low level SSB driving signal. If the modulation takes place in the high level RF amplifier, however, linearity is not a problem. High audio power is required for high level modulation, but today's digital techniques make possible pulse-width modulated audio systems having a 90 per cent, or better efficiency level.

Finally, in some UHF frequency control systems a subaudible tone must be transmitted to phase-lock the product detector of the receiver. A typical SSB crystal or mechanical filter does not pass frequencies closer to the carrier than 250 to 300 Hz without sacrificing sideband rejection. But with a phasing transmitter, there's no such problem, because there's no low frequency limitation to the audio signal transmitted. The system can handle modulation down to the DC level.

The modern UHF SSB transmitter makes use of high-power FETs because of their high class C efficiency. A block diagram of part of the audio phase shift and modulation system is shown in **fig. 1**. The audio signal is amplified and compressed and run through a bandpass filter. In the following buffer stage, the pilot or control tones are added to the audio signal. It then passes through a conventional audio phase shift network which provides two ports, separated by 90 degrees in phase. A differential amplifier follows the network. The amplifier is temperature compensated by mounting the two units on a common thermal link so that stable and equal gain are achieved in the two chains.

The audio signals are fed to a voltage-controlled astable multivibrator which has a resting frequency of 100 kHz with a 50 percent duty cycle. This stage drives a monostable multivibrator with a 10 microsecond reset time. With no modulation, these devices produce a square wave of 100 kHz with a 50 percent duty cycle. Any change in the frequency of the voltage controlled multivibrator will result in the change in the duty cycle of the following stage.

The multivibrators drive switching transistors (Q1, Q2) that reproduce the driving signal at the appropriate power level. The output passes through an integrator (L,C) which provides an exact reproduction of the input audio signal applied to the system. A protective diode (CR1) suppresses switching spikes that might damage the switching transistors.

The push-pull audio signals are fed to high level balanced modulators whose carriers are shifted 90 degrees in phase. The result represents cancellation of one sideband and reinforcement of the other. Upper or lower sideband can be chosen by reversing the audio phasing.

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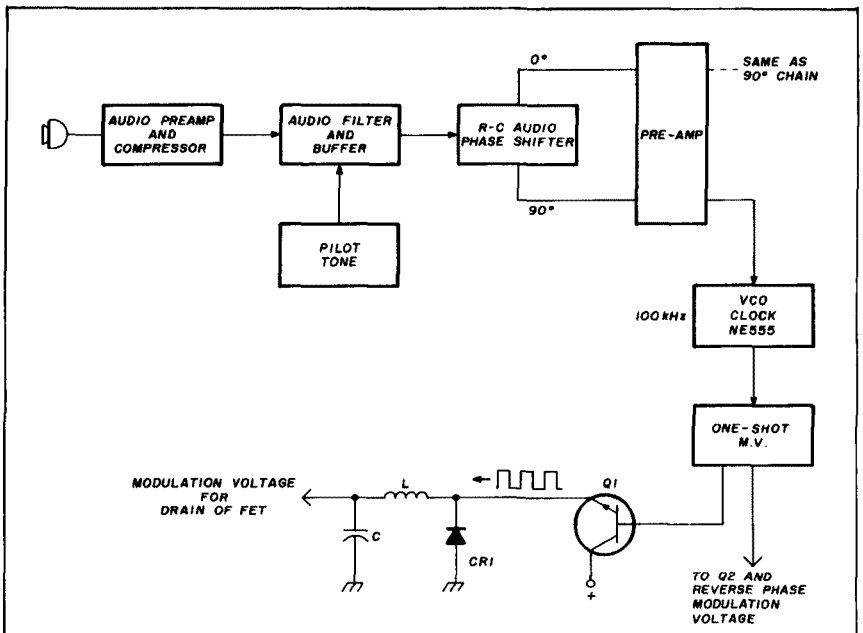


fig. 1. Pulse-width modulator for drain modulation of FET. Q1 is a high current switching transistor.

For high frequency operation, the RF phase splitter employs conventional coil and capacitor circuits. However, in the UHF region, these are replaced by a stripline hybrid circuit for stability and simplicity.

So the phasing system rises from the ashes to take part in today's exploration and utilization of the UHF frontier. Amateurs will soon rediscover this old technique and put it to new uses.†

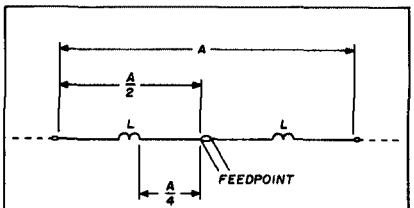


fig. 2. Coil-loaded dipole antennas. See *fig. 3* for coil data.

the compact dipole: the general case

In previous articles I've discussed coil-loaded, compact dipoles of interest to the Amateur who wants to operate on the HF bands but has limited antenna space. The designs shown, which placed the loading coil in the middle of the dipole leg, represent specific instances of the general case, where the loading coil (L) may be placed at any point in the dipole leg (fig. 2).

There are good reasons why the loading coil is centrally located in a dipole element. If the coil is placed near the feedpoint, less inductance is required, but current in the coil is higher and losses are higher. If the coil is located near the dipole end, less current flows in the coil, but more turns are required, increasing coil loss. Theoretically, if the coil were to be placed at the end of the dipole, the number of turns required would be infinite. (Efficiency aside, valid reasons may exist for placing the loading coil at some point in the dipole other than the center of the leg. It has been proven that feedpoint loading works well, even though loading coil efficiency may not be the highest — but that's another story.)

A good summary of dipole loading is contained in *The ARRL Antenna Book*,[1] in which a chart provides general design information for dipoles of various length and loading.

For the case of the center-loaded leg, the ARRL chart may be simplified as shown in **fig. 3** to indicate the reactance (X_L) required in a loading coil for antenna sizes ranging from 10 to 80 percent of normal. The graph is quite linear in the center portion but departs from linearity at both extremities.

As an example, assume a dipole only 100 feet long is needed for 3.8 MHz operation. A normal-sized dipole (made of wire) at this frequency would be 123.16 feet (37.53 meters) long. The required dipole, then, would be 100/123.16 = 0.81, or 81 percent the length of the regular dipole. This point falls just outside the right-hand edge of the graph, but the required loading coil reactance can be estimated as about 305 ohms.

The inductance required is found by the formula:

$$L = \frac{X_L}{6.28 \cdot f}$$

where **f** is the frequency in MHz and **L** is the inductance in microhenries.

A second example: assume a dipole is needed for 1.8 MHz operation, but can be only 40 percent the length of a full-size dipole, which would be 260 feet. The compact dipole length would be 260 x 0.4 = 104 feet (31.69 meters) long. Finding 40 percent on the x-axis and tracing the point up to the curve and over to the y-axis indicates that the loading coil for this antenna should have a reactance of 1,300 ohms. The formula given above will convert this value into microhenries.

Enterprising Amateurs have taken this whole concept and computerized it into a simple program so that once the antenna size is established in relation to a full-size dipole, the inductance of the coil, the number of turns, and coil diameter can be quickly determined for a center loaded element. How about that, computer fans?

† This material prepared and abstracted from "Technology Rejuvenates the Phasing SSB System," by Eric P. Nichols, in the August, 1985, issue of *Microwaves & RF* magazine, a Hayden Publishing Company monthly publication. (Hayden Publishing Company, Inc., 10 Mulholland Drive, Hasbrouck Heights, New Jersey 07604).

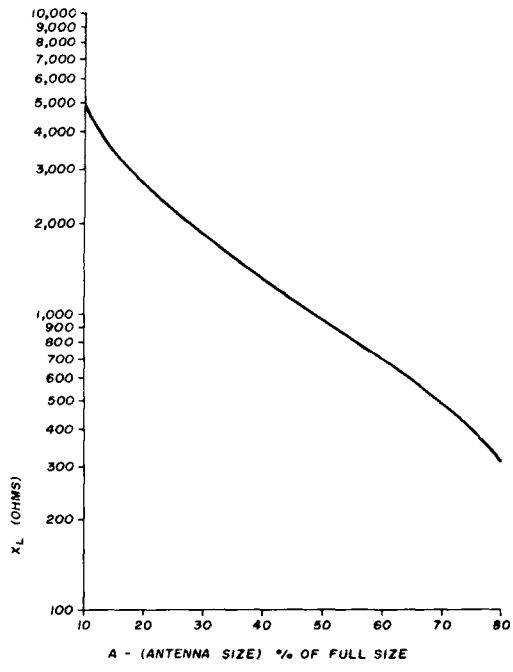


fig. 3. Loading reactance as a function of antenna length (*A*).

and now the bad news

Every HF operator is thoroughly familiar with the "Russian Woodpecker," the high power OTHB (over-the-horizon, backscatter) radars run by the Soviet Union. Repeated complaints about this obnoxious QRM generator have gone unheeded over the years. And now, I understand, the United States will soon have one running. But tests conducted with an experimental American OTHB radar revealed that the final version will not be a QRM-blaster as the Russian one has proven to be.

The American OTHB was discussed in some detail by O.G. Villard, Jr., W6QYT, writing in *QST*.[2] And the April, 1985, issue of *Amateur Radio*, the publication of the Wireless Institute of Australia, revealed that the prototype of an Australian OTHB radar was built and tested in 1978. The radar is located in the general area of Alice Springs, with the actual transmitter some 60 miles to the northeast. It is reported that the installation has successfully tracked commercial aircraft flying along the air route from Singapore to Australia. The radar receiving antenna array, which includes 468 pairs of broadband monopole antennas with each pair phased and sitting atop a very large ground mat, stretches over a distance of 1.7 miles (2.8 km). (*That should be very effective in a DX contest.*)

The power output of the transmitter is not disclosed, but it is reported to be over 400 kW. An array of vertically polarized log periodic antennas are used, with the beam steerable in azimuth.

Received echo signals, which are very weak, are analyzed in a computer. Frequency shifts on the returned signals caused by moving targets are analyzed to give an indication of the target speed. Thus the return signal from an aircraft is easily distinguished from a return signal from a small ship or other surface vehicle.

A frequency surveillance system prevents transmitted signal from operating on any occupied channel. The author of the article, Ian Hunt, of the Australian Department of Defense, writes that he "can state with certainty that the HF Amateur Radio bands are thoroughly protected in this manner. This fact...should make it obvious that statements (claiming) that OTHR signals emanating from the Jindalee system are observed in the Amateur bands would be ill advised, alarmist, and irresponsible in nature."

The OTHR system can track commercial aircraft for safety, keep track of military flights, detect unauthorized flights (such as used in drug smuggling), and watch wave systems on the ocean to advise shipping of possible hazards. It can also be used for increasing the knowledge of the ionosphere and the study of long-haul propagation.

how does the OTHR affect ham radio?

The Australian OTHR appears to be a well-run project that promises to produce little interference to the various high frequency services, including Amateur Radio. It's comforting to know that this radar is programmed to avoid the Amateur bands. Good thinking!

Pandora's box?

The photographs of the Australian installations reveal the immense size of the project. Yet dollar-wise, it seems to be quite modest. It seems to me that Australia, unintentionally, may be pointing the way toward the day when any country with a few bucks and lots of open space can set up its own "do-it-yourself" backscatter radar system. The newcomers to the game may not be so charitably inclined as our Australian friends, and in a few years we could all be plagued by many worldwide backscatter radars, situated all over the globe and making HF communications a thing of the past.

references

1. *The ARRL Antenna Book*, 14th edition, American Radio Relay League, Newington, Connecticut 06111, page 10-5, (figs. 9 and 10).
2. O.G. Villard, Jr., W6QYT, "Over-the-Horizon or Ionospheric Radar," *QST*, April, 1980, page 39.

ham radio

broadband RF transformers

Add fractional turns ratio techniques to your impedance matching arsenal

Despite conventional wisdom, which holds that a transmission line transformer can be built only with integral turns ratios, it's entirely possible to build a practical 2:1 impedance change transformer — that is, a 1.5:1 turns ratio transformer. Many other fractions are also possible. But first let's review the evolution of RF transformers, keeping an eye out for some of the possible pitfalls along the way.

basic transformer

Consider a basic autotransformer, which has one winding, with a tap. If there are five turns from start to tap, and two more from tap to finish, then the input impedance from start to finish, assuming 25 ohms between start and tap, is (25 x 7 x 7)/(5 x 5), or 49 ohms. This is within 2 percent of 50 ohms, so it could serve as a 2:1 transformer.

leakage inductance

Leakage inductance occurs when the flux from any part of a turn doesn't cut all of another turn. Leakage inductance is always present — it's just a question of *how much*. One might think that any flux that cuts the cores couldn't possibly contribute to leakage inductance, because it would then couple to the other turns. Thus many early transformer designs stressed placing the windings close to the cores. Note that by definition, both the primary and secondary have separate, and therefore probably unequal, leakage inductances. These leakage inductances limit the high frequency response because they are very real inductors in series with the transformer's leads.

twisted and braided windings

One might think that lowest leakage inductance is achieved when all the wires are wound tightly together. Imagine a rope consisting of five parallel strands of insulated wire twisted together. This rope is wound through a core for five turns only. Then a group of three of the strands are unwound one turn and then cut off, leaving two strands with five turns and three strands with four turns. The three four-turn strands are then connected in parallel to one another. These are in turn connected in series, aiding with both of the remaining five-turn windings. To summarize: from the bottom up, there are five turns in series with five turns to the tap, and then four turns to the top-end, producing a 14:10 ratio.

Another idea is to use seven strands twisted together as a rope with only one turn of this rope around the core. Connect five of these turns in series out to a tap, then add the remaining two in series. In theory this method might offer the lowest leakage inductance of any construction, but I can't handle the rat-nest of splices and still keep the leads short! And both of these techniques ignore capacitive effects that limit high frequency response.

transmission line

As the primary and secondary wires are brought closer to each other, rather than closer to the core, their mutual inductance and capacitance increases. It then becomes easier to envision the pair of wires as a transmission line. At first one might think the extra capacitance would limit the high frequency response, but it doesn't. The capacitance and leakage inductance combine to make a transmission line. The closer the wires are, the lower the transmission line's impedance. Considering the windings as transmission lines is a useful conceptual aid. It is now the standard method for making RF transformers, because it leads to wider-bandwidth performance. The high frequency end is helped without affecting the low frequencies.

bandwidth limitations

A transformer's low frequency response is a function of the type of core, its cross-sectional area, and the number of turns. The high frequency response depends on the inductance and the turn-to-turn capacitance as well as the length of the winding. The limiting length occurs when the induced voltage of a single turn adds to the next turns' (voltage) at a significantly *later* angle. This equates to a transmission line whose delay angle is determined by its length and velocity of propagation. The phase delay of the transformer limits its high frequency usefulness.

balanced transmission lines

Consider a pair of parallel conductors arranged so closely that when the current in one conductor flows in one direction, the current in the other flows in the

By Will Herzog, K2LB, 406 Yarmouth Road, Rochester, New York 14610

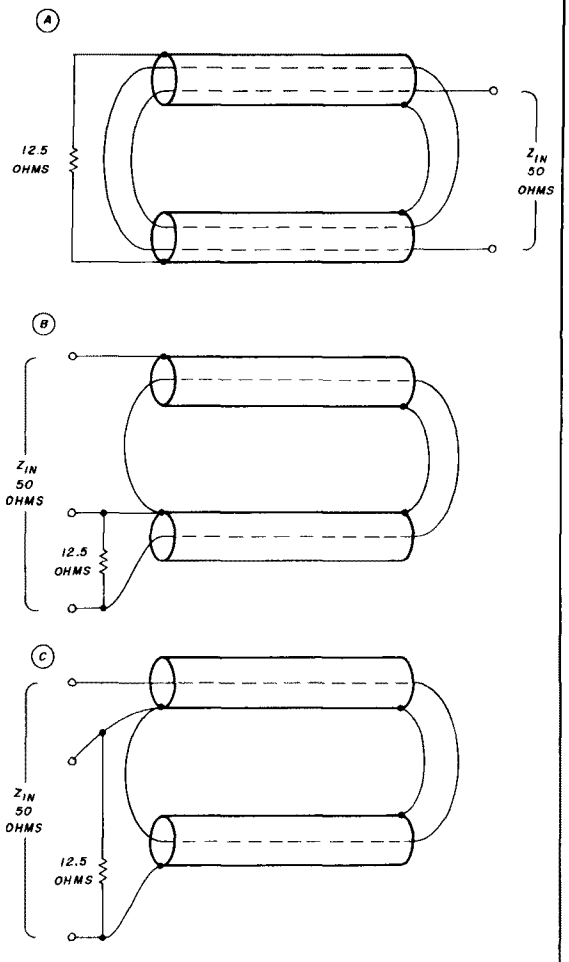


fig. 1. 4:1 impedance change tubing transformers: (*A*), connected as a transformer; (*B*) and (*C*), connected as an auto-transformer.

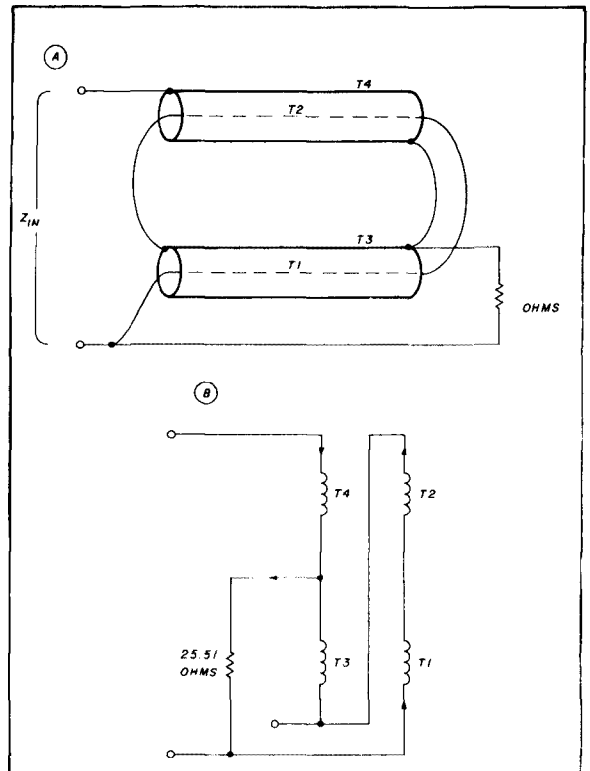


fig. 2. (*A*) Pictorial and (*B*) schematic representation of a tubing transformer connected to provide a 2:1 impedance change. (The reason it *won't* work is explained in the text.)

opposite direction. (Both are equal in magnitude.) This allows their magnetic fields to cancel out as soon as one is outside the space between the wires. As long as the currents are equal, the magnetic field does not need to intercept the cores. The inter-wire space also has an electric field due to the potential difference between the wires. The electric field diminishes as soon as one is away from the wires several times their spacing, especially if a material with a high dielectric constant is between the wires. Consequently, electromagnetic field propagates between the wires and transmission line theory applies. The uncoupled magnetic lines result in a leakage inductance, analogous to a transmission line unbalance which results in line radiation.

transmission line impedance

The impedance of a transmission line is sometimes defined as the square root of the ratio of inductance per unit length divided by the capacitance per unit length. If one has inductance and capacitance meters, this can be measured easily by shorting the output end of the transmission line and measuring the inductance at the other end. Capacitance is measured directly after the short is removed. Measurements are taken at a frequency in which the line length is less than 1/10 of a wavelength. Divide the Henries by the Farads and take the square-root of this ratio to get the impedance of the transmission line in ohms. A vector impedance meter can be used to measure the line impedance by varying the frequency at about the 1/8 wavelength point until the magnitude of impedance stays the same as the other end of the line is alternately shorted and left open. The phase swings from inductive to capacitive; its amount is ignored. The surge impedance can be read directly.

We must be able to identify every conductor with a mate carrying current in the opposite direction, resulting in the absence of an external magnetic field and consequently small leakage inductance. However, even this procedure is difficult, since the capacity of each wire is not only to every adjacent wire(s), but also to ground; and the currents in parallel conductors still

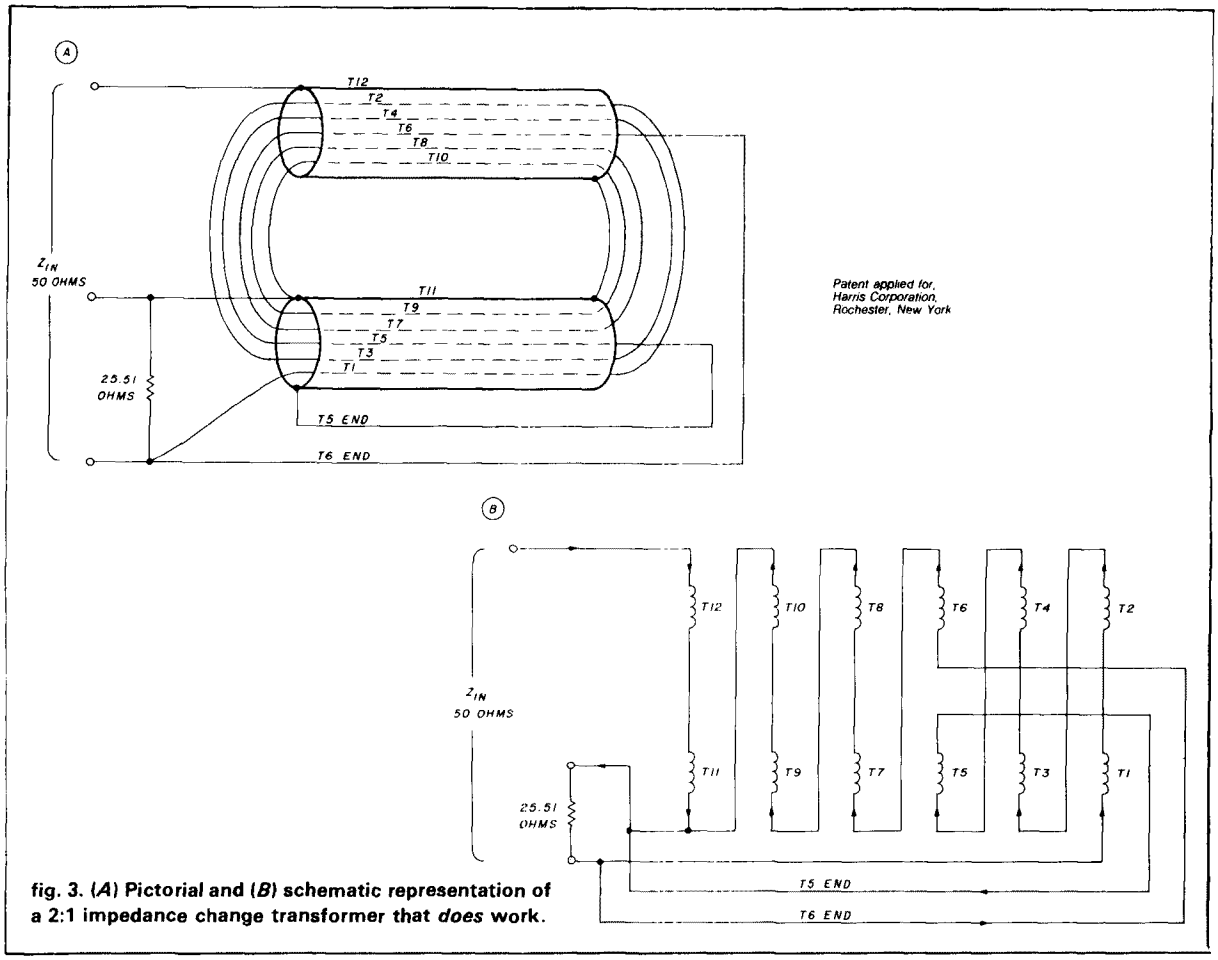


fig. 3. (*A*) Pictorial and (*B*) schematic representation of a 2:1 impedance change transformer that *does* work.

must be kept equal and opposite for each subgrouping of closely spaced conductors.

This means that every wire can have several individual characteristic impedances, each of which should be constructed to provide the same impedance as obtained by dividing voltage by current, or Z_ϕ. Z_ϕ is determined by wire size, spacing, insulation, and number of conductors. The ultimate limit in high frequency response is the physical length of all the turns, considered as a percentage of its electrical wave length. There's another paradox here: if one looks at it as a conventional transformer, the induced voltage of each turn must be the same because the flux is changing at the same time in each turn. However, with transmission line analysis, there isn't any external field, so it's much easier to see that it's the total length that counts.

tubing transformers

Tubing transformers, now very popular in transistor transmitters, use two parallel sections of metal tubing electrically connected at one end. The free ends of the tubing have terminals. The tubing itself can then be used as part or all of one of the windings, or the tubing may simply be grounded to serve as an electrostatic shield for the cores. Cores of ferrite or powdered iron are threaded onto the exterior of the tubes to increase the low frequency inductance. Wire is then run through the interior of the tubing to make the other winding(s). The wires can come out at either the terminal ends of the tubing or at the common connected end; it makes no difference. The wires must make a complete turn, so the turns ratio is again always an integer. **Figure 1A** shows a 12.5 to 50-ohm transformer. **Figures 1B** and **1C** show a similar construction, connected as an autotransformer to transform 12.5 to 50 ohms.

If a transformer similar to the one shown in **fig. 1C** were used to transform a 25-ohm resistor to 50 ohms by tapping up on the winding or tubing to get a fractional turn, the results would be poor (see **fig. 2A.**) In **fig. 2B** arrows indicate the direction of the currents. Note that T2's and T4's currents are in opposite directions canceling the magnetic fields. But T1's and T3's currents are in the same *direction*! This means that the fields can't cancel, so instead of a transmission

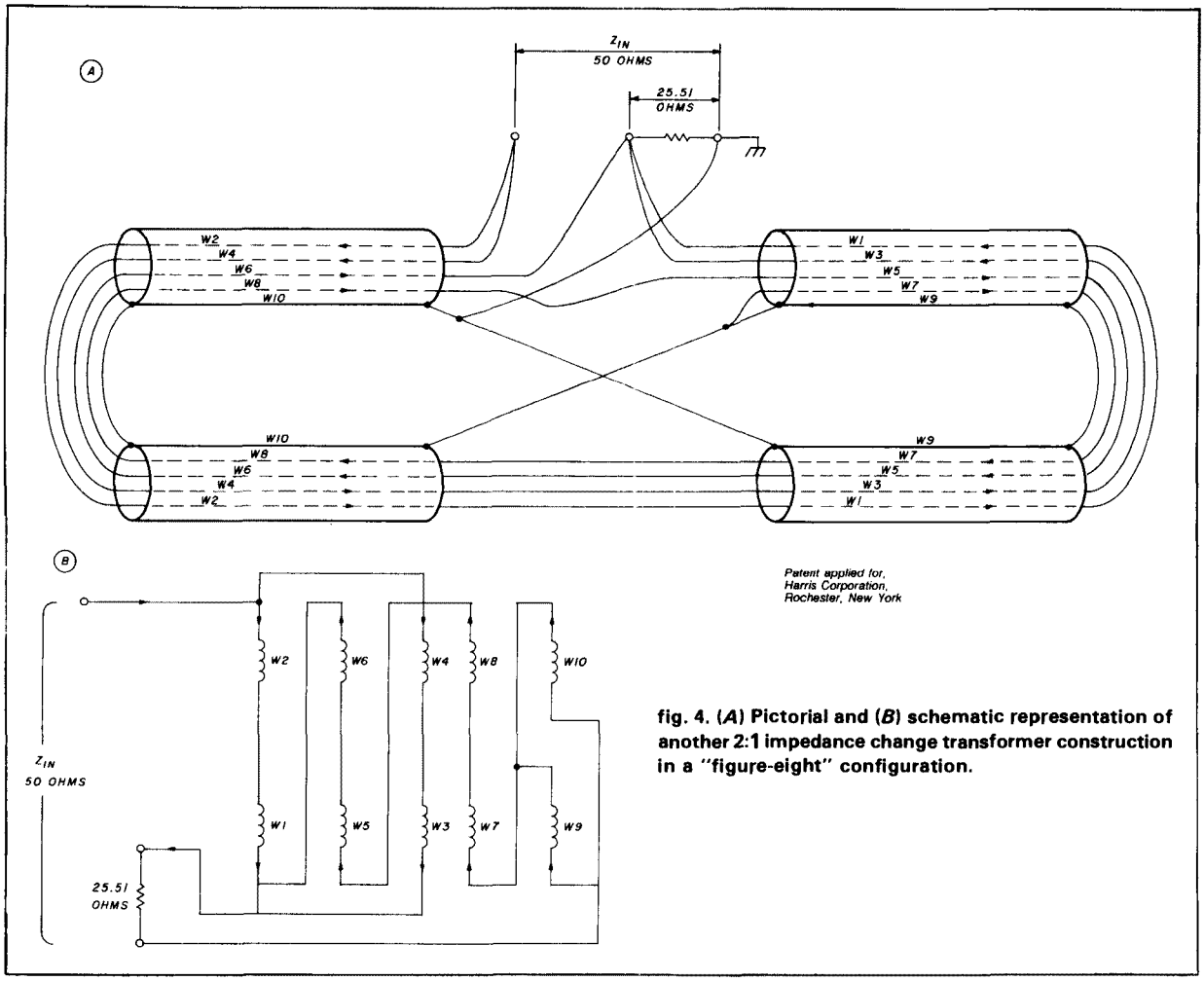


fig. 4. (*A*) Pictorial and (*B*) schematic representation of another 2:1 impedance change transformer construction in a "figure-eight" configuration.

line, we have a very good inductor — just what we *don't* need because it limits the high frequency response.

split the winding

Yes — you can use fractional turns and still have a transmission-line type device. The key is to add an extra winding that's mostly in parallel with the first so that two separate windings exist. The first winding has an extra half turn; the second winding skips that half turn and picks up the remaining half turn. Then the pair will function just like a single winding with a half-turn, but without a flux or current unbalance! Each of the two windings carries half as much current as the single winding did before it was split into two windings, so they can be made of smaller wire. (See figs. **3A** and **3B**.)

The fractional turn transformer in **fig. 3** has a turns ratio of 7:5, which gives an impedance ratio of 49:25. The secondary has two windings in parallel: T1 through T5, in five sections; and T6 through T10, also in five sections. The primary has seven sections because it adds T11 and T12 to the secondary's five sections. Thus if one counts the complete turns, it's a three-and-a-half-turn to two-and-a-half-turn transformer! Note that the T5 end and T6 end are longer leads, but that their currents flow in opposite directions, and the voltage between them is the secondary voltage; thus they can be envisioned as a transmission line. The fields are balanced in one direction because the currents are equal/and opposite, so the net ampere-turns is zero. It works well.

ampere-turn and Z_{ϕ} analysis

To see this balancing action more clearly, consider that for the amp-turns to be zero, the current total of T2, T4, T6, T8, and T10 must equal T12's. Let's assume 25 volts across the secondary load resistor, so the output current is then 1 ampere. The power out is (25x25)/25.51 = 24.5 watts. From symmetry, the voltage across each of the "T" sections must be equal to each other, and since there are five in series across the secondary, each section has 5 volts. The total voltage input across the seven sections is

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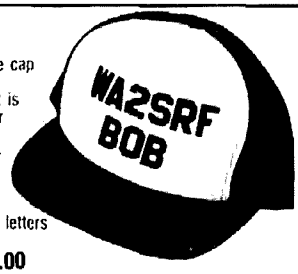
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then 35 volts. If we assume a lossless transformer, there must be 24.5 watts input. The input impedance is, then, equal to (35)(35)/24.51 = 50 ohms. The input current in T12 and T11 must therefore, from Ohms law, equal (35 volts)/(50 ohms) = 0.7 amperes. Likewise, the current out of the transformer is (25 volts)/(25.51 ohms) = 0.98 amperes. Since T5 and T6 are in parallel and symmetrical, the current flowing through each of them is equal and is: I = (I out—I in)/2 or (0.98—0.7) = 0.14 amperes.

Because we know all the currents, we can now calculate the amp-turn balance. Again, refer to the direction of the arrows in **fig. 3B**, noting that T2 plus T4, plus T6, plus T8 plus T10's currents must equal T12's. Five times the 0.14 ampere = 0.7 ampere, which was T12's current, verifying that the net external flux is zero.

The impedance of the T5-T6 lead transmission line should be 25 volts divided by 0.14 ampere or 178.6 ohms. Spacing the leads in the air can achieve this.

heat problems

Note that the tubing, which is the largest conductor, is used for the section of the transformer that carries the most current. This is good news. The bad news is that if high power operation is required, one finds that the hot spot is the ferrite cores, not the windings, which heat because of dielectric loss in the cores. The electric field is high on the cores because the tubing is at the highest RF voltage and the transformer is mounted near a chassis ground. Thus another novel technique, resulting in longer life at high power, evolved.

A tubular transformer, shaped as a figure-eight, was developed. This transformer allowed placing the leads closer together, eliminating the external line section, and permitting the tubing to work both as a better electrostatic shield and as the half-turn winding, thereby allowing the tubing to be placed near the chassis ground potential. The highest potential between the tubing and the cores is only a half-turn's worth, because it is now also used as the split turn. Note that in **fig. 4**, two labels are provided on each wire. These two quarter-turns are hereafter referred to as one half-turn. The turns of the secondary are W9 in parallel with W10, which serve as the half-turn, plus W7 and W8, W5 and W6. This makes a two-and-a-half-turn secondary. The primary totals three-and-a-half-turns because it adds to the secondary W3 plus W4 in parallel with W1 plus W2. The top primary turns are split into two paralleled sections because it carries the most current, and because physically W1-W2 and W5-W6 can be close-spaced to make one transmission line, and W3-W4 and W7-W8 are close-spaced to make a second transmission line.

ham radio

NBS trigon
is improved
by increasing
element lengths

optimizing
a Yagi trigon reflector

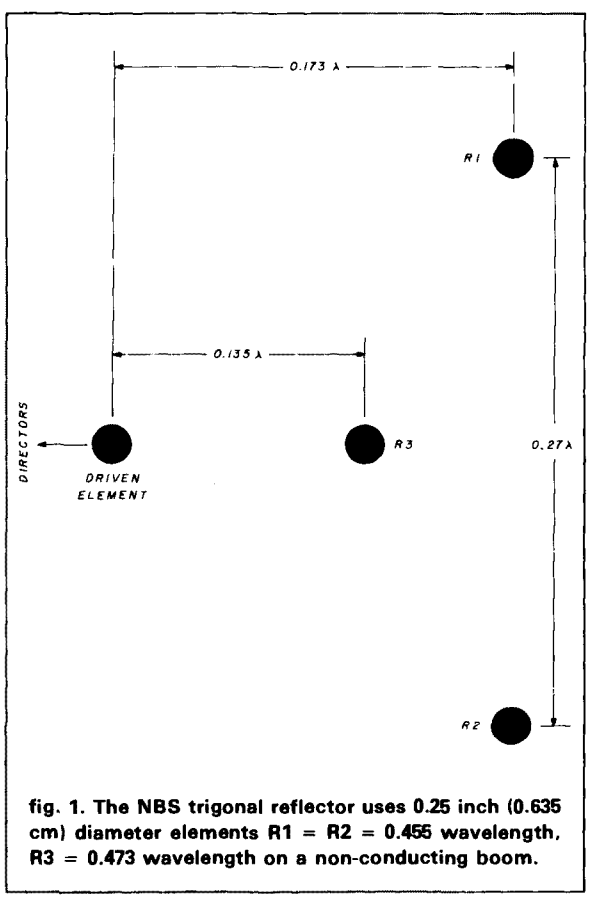


fig. 1. The NBS trigonal reflector uses 0.25 inch (0.635 cm) diameter elements R1 = R2 = 0.455 wavelength, R3 = 0.473 wavelength on a non-conducting boom.

One of the most valuable sources of Yagi antenna designs is *NBS Technical Note 688*.[1] In addition to the seven measured and patterned Yagis fully discussed, it also includes suggestions for further enhancing Yagi designs.

One of the suggestions presented is the use of a three-element (trigon) reflector to increase the gain of the 4.2 wavelength Yagi. The trigon provided a stated 0.75 dB gain increase for this Yagi, representing the maximum increase obtainable by using more than one reflector. The 4.2 wavelength Yagi was the only Yagi with which this type of experiment was performed.[2]

Because the possibility of obtaining more gain from the six shorter Yagis was intriguing, it was pursued with the same Yagi computer model I described in several previous *ham radio* articles.[3] But analyzing the use of parasitic elements not in the same plane with other Yagi elements would have required extensive modification to the model's code; to avoid that chore, I chose to use other software capable of modeling elements in multiple planes.

At the 1983 and 1984 Dayton Hamventions, Dr. James Breakall, WA3FET/6, described his use of the Numerical Electromagnetics Code (NEC) as the basis for his presentations on Yagis, dipoles, and vertical antennas. A complex FORTRAN software system, NEC requires a mainframe for execution. A shorter version, MININEC, is written in BASIC and is designed specifically for execution on microcomputers.* For those familiar with MININEC, the entire NBS series of Yagis was modeled using both eight and 18 segments per element. However, to speed the modeling process, trigon Yagis were stated in terms *of* eight segments per element.

By Stanley Jaffin, WB3BGU, 800 Stonington Road, Silver Spring, Maryland 20902

*An Apple-readable version of the MININEC code can be obtained by sending a blank floppy disk to: Dr. Edmund K. Miller, L-153, Lawrence Livermore National Laboratories, P.O. Box 5504, Livermore, California 94550.

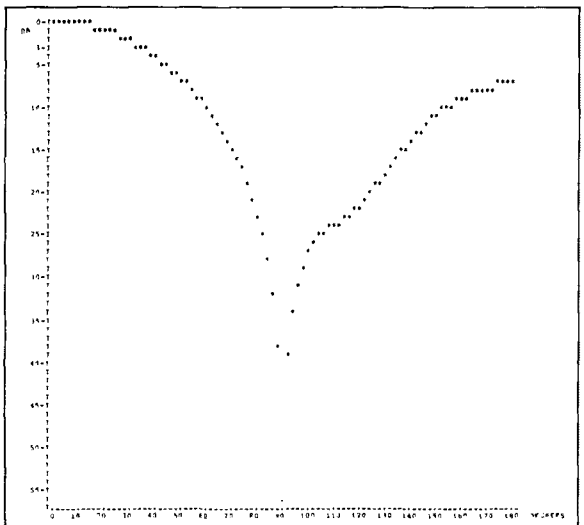


fig. 2. E-plane plot of NBS 0.2 wavelength, 2-element Yagi measured at 400 MHz. Notice in this and in *figs. 3* through *15* the pattern amplitudes are normalized. Maximum response occurs at boresight (0 degrees) and is labeled 0 dB.

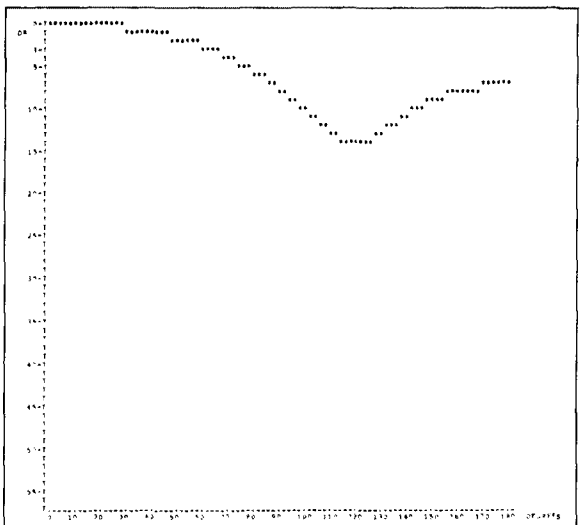


fig. 3. H-plane plot of NBS 0.2 wavelength, 2-element Yagi measured at 400 MHz. This Yagi shows a calculated gain of 6.62 dBi and a F/B of 7.28 dB.

Table 1 contains a summarized comparison of the published NBS findings, the WB3BGU model, MININEC, and pattern integration based on the radiation patterns in reference 1.[4] The differences in gain figures are due to the different pattern integration code executed in each of the two models. Even slight variations in the overall pattern can affect how a single lobe (the main lobe) is defined in terms of the overall pattern. An examination of the seven E-plane and seven H-plane patterns resulting from each method of analysis (not shown) indicates a high degree of qualitative agreement. In many instances the MININEC patterns were more indicative of nuances in the NBS plots than the WB3BGU model. Reference 1 states that the original gain figures are accurate to within 0.5 dB, meaning that the NBS baseline is not exact. This is in part due to the fact that the NBS Yagis were measured for gain on one range but patterned on another.[5] The lack of pattern symmetry indicates that the range

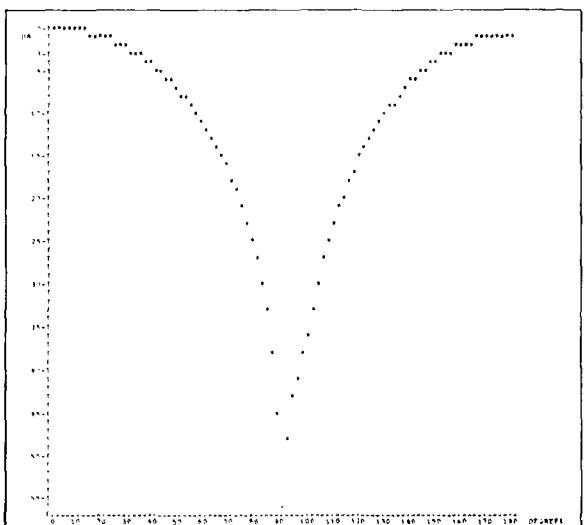


fig. 4. E-plane plot of baseline (stock) trigon reflector and driven element Yagi measured at 400 MHz.

table 1. Comparative gain and F/B figures for the NBS Yagi family. Note: *NBS Technical Note 688* did not specify exact F/B measurements. The NBS F/B figures are taken from the cartesian plots provided for each Yagi.

NBS Yagi antenna		NBS		WB3BGU		MININEC-8		MININEC-18		pattern integration
boom in wavelengths	number of elements	gain (dBi)	F/B (dB)	gain (dBi)	F/B (dB)	gain (dBi)	F/B (dB)	gain (dBi)	F/B (dB)	gain (dBi)
0.2	2	6.71	6	6.70	6.71	6.62	7.28	6.58	7.90	6.71
0.4	3	9.15	8	9.15	12.07	9.27	11.82	9.47	11.97	9.62
0.8	5	11.35	15	10.71	22.71	10.69	24.53	10.93	20.76	11.41
1.2	6	12.35	19	11.75	19.05	11.67	19.11	11.97	27.88	12.64
2.2	12	14.40	23	14.00	30.47	13.75	22.68	13.99	36.43	14.28
3.2	17	15.55	24	15.22	24.06	14.99	22.55	15.20	27.82	15.47
4.2	15	16.35	22	15.70	19.26	15.38	21.21	15.72	21.93	16.22

used for patterning may not have been totally free of reflecting surfaces. Nevertheless, MININEC's calculated gain, F/B, and patterns can be considered to be very good replications of the NBS findings.

the NBS trigon reflector

Reference 1 provides a list of the multiple reflector arrangements NBS examined during the trigon evaluation. Maximum gain, the sole criterion in selecting the trigon configuration, is used as the main criterion in evaluating the modeled trigons. As is the case for all other NBS parameters, an element diameter of 0.25 inches, a design frequency of 400.0 MHz, and a nonconducting boom apply. The stock NBS trigon is shown in **fig. 1.**

The 4.2 wavelength Yagi was modeled with the stock trigon. A significant reduction in both gain and F/B resulted. The first experiment involved varying the trigon element lengths (spacing was held constant) in order to achieve maximum gain. All three trigon elements were iterated over a range of plus and minus 0.015 wavelengths, in increments of 0.001 wavelengths. Increased gain correlated with increased trigon element length, reaching a calculated maximum with simultaneous extensions of 0.009 wavelengths. These iterations were repeated for each of the remaining six Yagis. **Table 2** contains the results of replacing the 0.2 wavelength Yagi's single 0.482 wavelength reflector with the stock (0.000 increment) trigon and successively incremented trigons. For this Yagi only, the 0.007 increment produces the maximum calculated gain. **Figures 2** and **3** contain the E- and H-plane plots for the original single reflector 0.2 wavelength Yagi. **Figures 4** and **5** contain these same plots using the stock trigon, and **figs. 6** and **7** present these same plots for the gain maximizing 0.007 increment trigon. For comparison, **figs. 8** and **9** do likewise for the 0.008 increment trigon, resulting in a minimal reduction in calculated gain but an increase of 0.77 dB in F/B ratio.

Either of the two incremented 0.2 wavelength Yagi versions would be useful as a single HF (7 MHz in particular) array or as a building block in a VHF/UHF col-

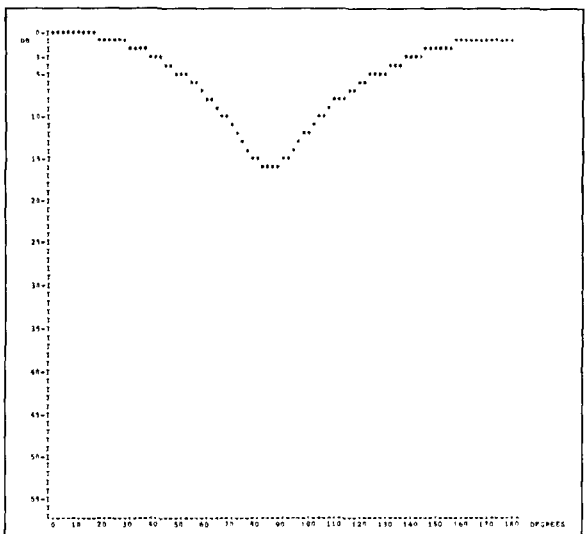


fig. 5. H-plane plot of baseline trigon reflector and driven element Yagi measured at 400 MHz. Gain is calculated at 6.42 dBi. F/B equals 0.76 dB.

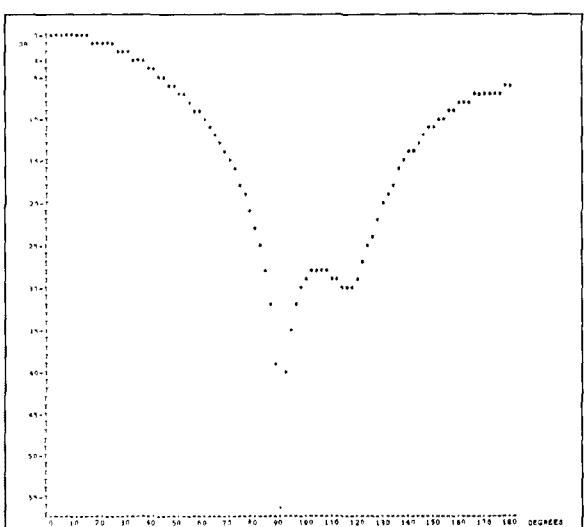


fig. 6. E-plane plot of 0.007 wavelength extended length trigon reflector and driven element Yagi measured at 400 MHz.

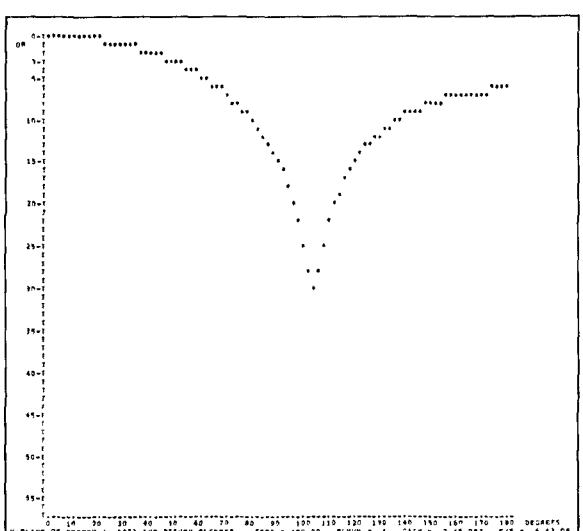


fig. 7. H-plane plot of 0.007 wavelength extended length trigon reflector and driven element Yagi measured at 400 MHz. Calculated gain equals 7.45 dBi and F/B equals 6.43 dB.

table 2. Gain and F/B calculations resulting from iterating the trigon reflector on the 0.2 wavelength NBS Yagi.

increment in wavelengths	gain (dBi)	F/B (dB)
0.000	6.42	0.76
0.001	6.54	0.99
0.002	6.69	1.33
0.003	6.87	1.85
0.004	7.07	2.64
0.005	7.26	3.75
0.006	7.40	5.10
0.007	7.45	6.43
0.008	7.44	7.34
0.009	7.40	7.66
0.010	7.34	7.58
0.011	7.29	7.32
0.012	7.25	7.03
0.013	7.22	6.79
0.014	7.20	6.59
0.015	7.18	6.45

linear array. At 7 MHz a relatively short 25-foot boom would yield a high gain Yagi with good sidelobe structure. At VHF/UHF this same array serves as a relatively high performance building block for collinear arrays. In table 3 the seven NBS Yagis are compared in terms of their respective single reflector versions and three trigon variations: stock, gain maximized, and F/B maximized. With the exception of the 0.2 wavelength Yagi, maximum gain occurs when each trigon element is extended 0.009 wavelengths from its stock length. With the exception of the 0.4 wavelength Yagi, calculated F/B maximas occur at even longer extensions. The 0.4 wavelength Yagi achieved both calculated maximas with the 0.009 extensions, and had the only single frequency F/B maxima.

table 3. Comparative performance calculations for the NBS Yagis with three specified sets of trigon reflectors. Note: Column 2 includes two additional elements in the trigon reflector. Column 1 shows the original Yagi length.

NBS Yagi antenna						gain maximum trigon		F/B maximum trigon		F/B maximum trigon
boom in wavelengths	number of elements	Basic Yagi gain (dBi)	F/B (dB)	stock trigon gain (dBi)	F/B (dB)	gain (dBi)	F/B (dB)	gain (dBi)	F/B (dB)	increment (dBi)
0.2	4	6.62	7.28	6.42	0.76	7.45	6.43	7.39	7.66	0.009
0.4	5	9.27	12.07	6.57	4.17	9.28	30.91	9.28	30.91	0.009
0.8	7	10.69	24.53	8.79	7.15	10.82	17.01	10.62	25.53	0.014
1.2	8	11.67	19.11	10.18	6.03	11.92	12.39	11.73	14.73	0.015
2.2	14	13.75	22.68	12.05	8.55	13.90	16.34	13.81	18.91	0.013
3.2	19	14.99	22.55	13.44	9.69	15.16	18.23	15.10	19.70	0.013
4.2	17	15.38	21.21	13.86	9.27	15.67	18.53	15.63	19.44	0.011

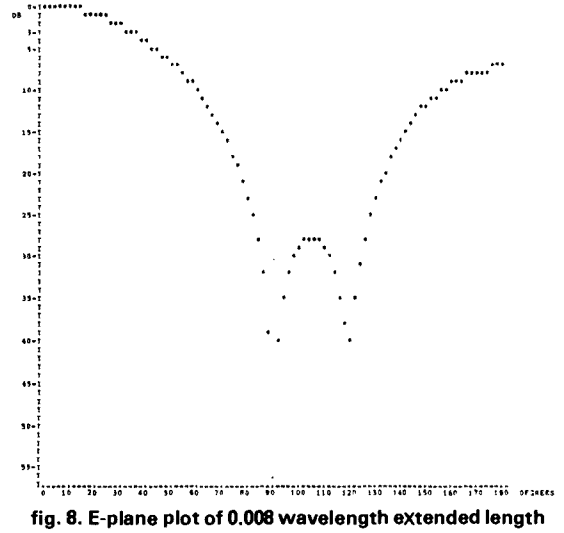


fig. 8. E-plane plot of 0.008 wavelength extended length trigon reflector and driven element Yagi measured at 400 MHz.

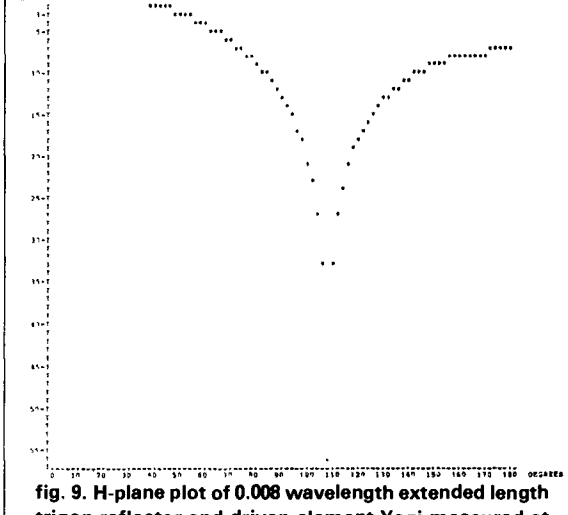


fig. 9. H-plane plot of 0.008 wavelength extended length trigon reflector and driven element Yagi measured at 400 MHz. Calculated gain equals 7.44 dBi and F/B equals 7.34 dB.

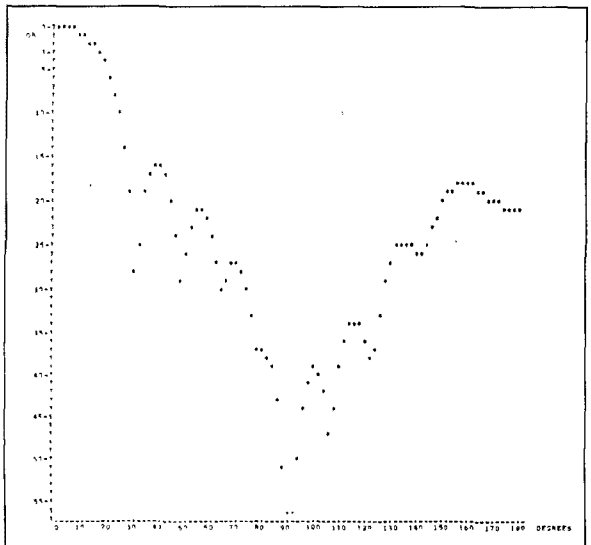


fig. 10. E-plane plot of 4.2 wavelength, 15-element Yagi that uses a single reflector.

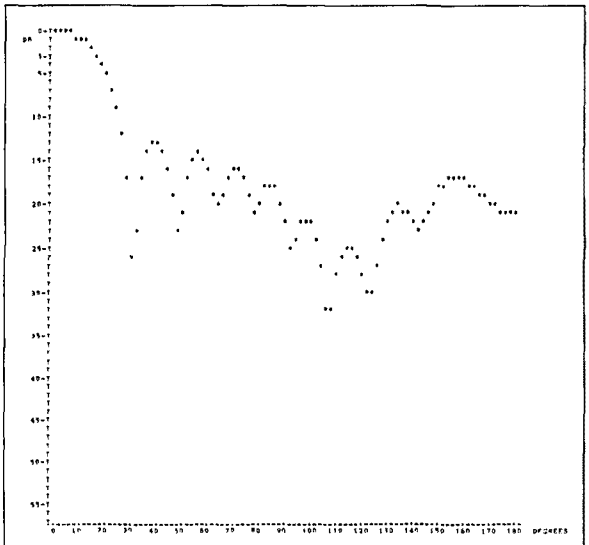


fig. 11. H-plane plot of 4.2 wavelength, 15-element Yagi that uses a single reflector. Calculated gain equals 15.38 dBi and F/B equals 21.21 dB.

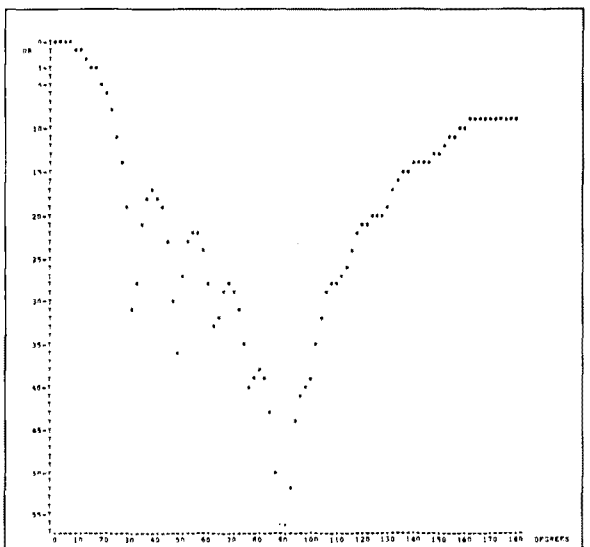


fig. 12. E-plane plot of 4.2 wavelength, 15-element Yagi that uses a stock trigon reflector.

Figures 10 and **11** illustrate E- and H-plane plots for the single reflector 4.2 wavelength Yagi. In **figs. 12** and **13** a stock trigon reflector is used. In **figs. 14** and **15** the 0.009-wavelength extended reflector length gain optimized version is plotted. For the first 80 degrees in the E-plane plot, the stock trigon actually improves the suppression of the minor lobes. This is also true for the first 55 degrees of the H-plane plot, but at the expense of gain, F/B, and other minor lobe levels in both planes. The gain maximizing 0.009 wavelength extensions provide the clean pattern normally associated with this Yagi, and additional 0.34 dB in gain, and further (1-2 dB) reduction of the first few minor lobes in both the E- and H-plane plots. While calculated F/B is reduced, the resulting first minor lobe permits an increased stacking distance and hence more incremental gain from a stack. As reference **1** did not report any patterns from the trigon experiment, it is not known whether NBS noted these other improvements along with increased forward gain.

validating the calculated trigon findings

Several questions about the validity of these calculated findings versus the published NBS findings arise. There is a measurable difference between the 4.2 wavelength trigon Yagi's calculated performance and what is given in reference 1. As NBS did not provide plots for this Yagi, even qualitative comparisons cannot be directly made. There is still the issue of the trigon's dimensions. Use of the stock element lengths results in degrading the 4.2 wavelength Yagi's performance as well as that of the six others.

Before these iterations were begun, Joe Reisert, W1JR, had performed extensive analyses based on the use of a trigon reflector of the 2.2 and 3.2 wavelength NBS Yagis on an antenna range.[6] He found that for these two Yagis the stock trigon reflector reduced the measured gain by just over 1.6 dB, and that a nominal gain increase of nearly 0.2 dB over the single reflector Yagi could be obtained by lengthening all three

reflector elements by approximately 0.008 wavelengths. Trigon element spacing remained constant. These empirical findings are in very good agreement with the MININEC calculations in **table 1**. Günter Hoch, DL6WU, has reported similar gain findings for a broad range of multiple element reflector arrangements.[7] While there is no reason to doubt that Peter Viezbicke, WØNXB, the author and principal investigator for the experiments reported in reference 1, actually measured what was reported, it is unfortunate that neither subsequent empirical experiments nor mathematical modeling can exactly reproduce these well known findings.

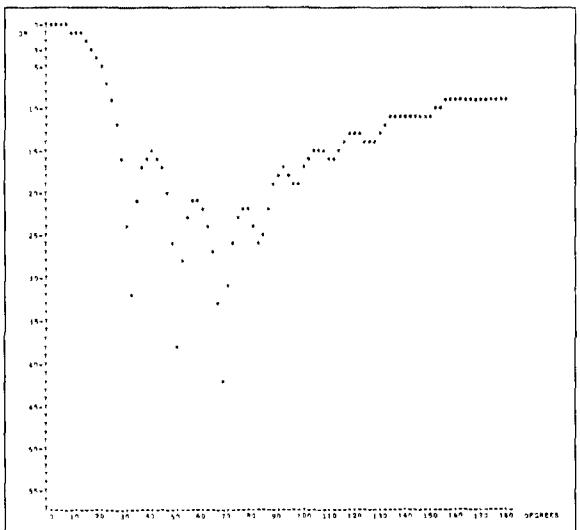


fig. 13. H-plane plot of a 4.2 wavelength, 15-element Yagi that uses a stock trigon reflector. The calculated gain is 13.86 dBi and the F/B equals 9.27 dB.

conclusions

The trigon reflector can provide increases in (calculated) gain for all seven NBS Yagis. For the 0.2 wavelength Yagi this increase is at its maximum value. For the 0.4 through 3.2 wavelength Yagis, this increase is mostly nominal. For the 4.2 wavelength Yagi used to develop the trigonal reflector, the increase is more pronounced, but less than initially reported. For all NBS Yagis it is necessary to use reflector element lengths (with spacing held constant) longer than the stock lengths reported in reference 1. The stock lengths severely reduce the performance level of any NBS Yagi on which they are used.

Reference 1 presents a series of design curves in its table 21 for converting director and reflector elements to other diameter-to-wavelength ratios. Mathematical analysis indicated that reflector curves for trigon elements that were drawn parallel to the curves for single reflectors would result in accurate conversions. As is the case with the other curves, these trigon curves had to be drawn through the point of intersection between 0.0085 wavelengths in diameter (X axis) and the trigon element's optimal length on the Y axis. This was confirmed by reference 6. A more recent and easier to read version of these curves had been found, and with the trigon curves is presented as **fig. 16.**[8] These curves can now be used exactly in the same fashion as the other design curves.

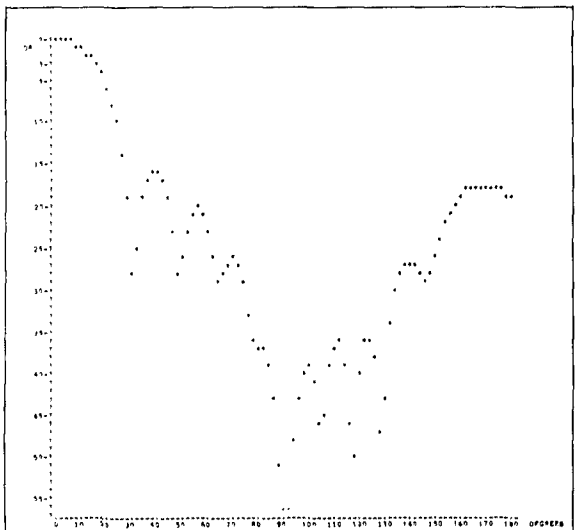


fig. 14. E-plane plot of a 4.2 wavelength, 15-element Yagi that uses gain optimized trigon reflector lengths.

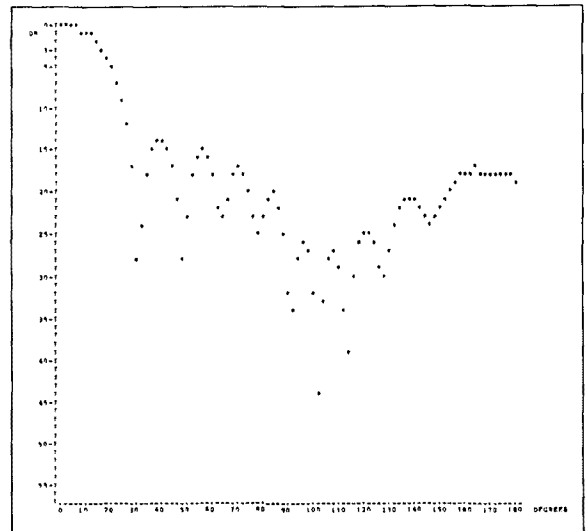


fig. 15. H-plane plot of a 4.2 wavelength, 15-element Yagi that uses gain optimized trigon reflector lengths. The calculated gain is 15.67 dBi and the F/B equals 18.53 dB.

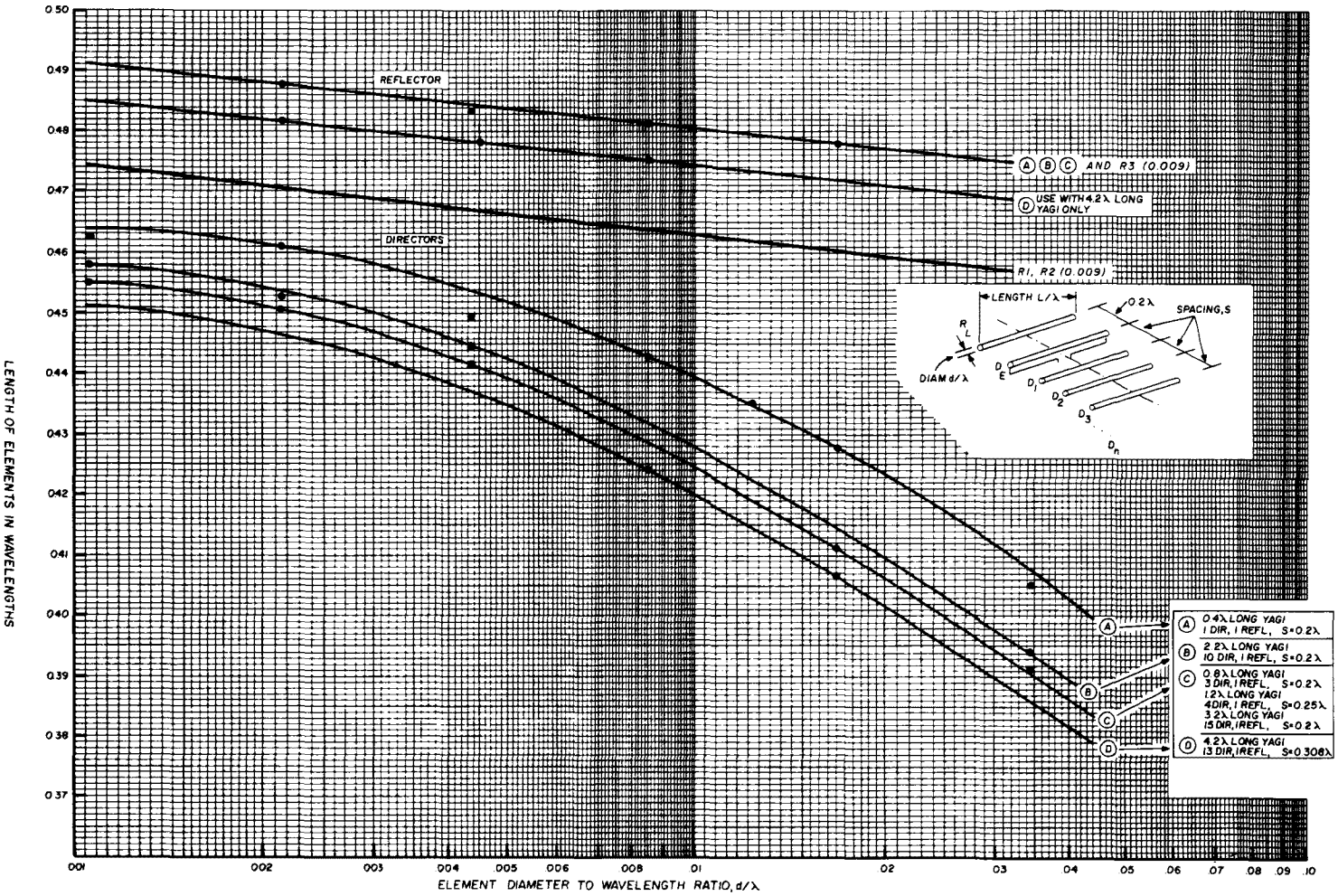


fig. 16. Yagi antenna design nomograph shows the relationship between element diameter and wavelength ratio (d/λ), and element length for different antennas.

The procedure for compensating for the diameter of a conducting boom is the same as for other NBS Yagi elements.

A great deal of experimentation can still be performed on the NBS Yagi antenna. Whether performed on an antenna range or by computer iteration, the results should be presented so that all Radio Amateurs can benefit from these experiments.

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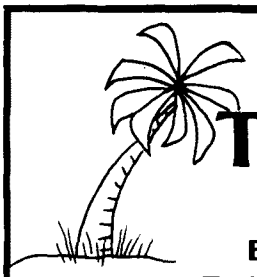
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Easy-to-build coax-stub filters help reduce both harmonics and intra-station interference

trapping stubborn harmonics

When you're trying to eliminate interference caused by harmonics, you have to stop the offending signals at the transmitter before they're radiated. In such multi-transmitter situations as Field Day and contests, harmonic interference can mean the difference between an interference-free operation and endless arguments over doing who's doing what to whom. By placing a coax stub filter at your transmitter's output, you can reduce harmonics at the source. What's more, coax-stub filters can also help reduce harmonic-induced TVI.

a simple solution

What makes coax-stub filters an easy solution to the problem is their simplicity; all you need is some coax and a few connectors. And stub filters don't have any coils to wind or capacitors to tweak. They don't require a sweep generator or network analyzer to tune. Just trim the stubs to length with the aid of a noise bridge.

Table 1 lists the HF Amateur bands and the filter combinations you can use to eliminate harmonics and, in the case of 20 meters and 15 meters, intra-band interference and front-end overload in your receiver. Each filter consists of one or two stubs.

how they work

A half-wave, single-stub filter acts as a subharmonic filter by introducing a short circuit into the transmission line at the frequency it's an open-circuited quarter-wave for. At the frequency for which the stub is a half wave, it looks like an open circuit and passes the desired signal.

A quarter-wave, single-stub filter acts as an even-harmonic filter by introducing a short circuit into the transmission line at the frequency it's a half-wavelength for. At the frequency for which the stub is a quarter wave, it looks like an open circuit and passes the desired signal. Because a shorted quarter-wave stub traps even harmonics, the 160-meter and 80-meter bands need just one filter each to trap all harmonics that fall in the Amateur bands below 30 MHz.

Some of the stubs introduce reactance on the line. In these cases, a second compensating stub tunes out the reactance by introducing an equal and opposite reactance. (The net effect is analogous to a parallel-resonant LC circuit; at resonance, the inductive and

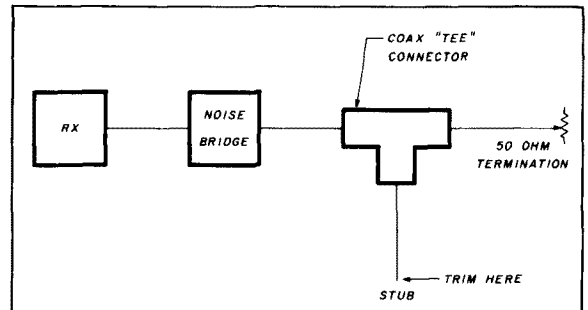


fig. 1. Trim a stub for 50-ohms resistance by moving a shorting pin down the coax until you get a null with the noise bridge set to 50 ohms and the receiver tuned to the band for which the stub is a quarter-wave. Use RG8 or equivalent in high-power applications.

By Robert M. Clarke, N1RC, 150 Stimson Street, West Roxbury, Massachusetts 02132

table 1. HF Amateur bands and filter combinations used to eliminate harmonics.

band	stub filter[1] needed to prevent interference to 160	80	40	20	16	10
160	x	$\frac{\lambda}{4}$ S	$\frac{\lambda}{4}$ S	$\frac{\lambda}{4}$ S	$\frac{\lambda}{4}$ S	$\frac{\lambda}{4}$ S
80	x	x	$\frac{\lambda}{4}$ S	$\frac{\lambda}{4}$ S	$\frac{\lambda}{4}$ S	$\frac{\lambda}{4}$ S
40	x	x	x	$\frac{\lambda}{4}$ S	$\frac{\lambda}{12}$ O[2] $\frac{\lambda}{6}$ O[4]	$\frac{\lambda}{4}$ S
20	x	x	x	x	$\frac{\lambda}{6}$ O[2] $\frac{\lambda}{12}$ S[4]	$\frac{\lambda}{4}$ S
15	x	x	x	$\frac{\lambda}{8}$ O[4] $\frac{3\lambda}{8}$ O[3]	x	x
10	x	x	x	x	x	x

S = short-circuited stub O = open-circuited stub

Notes:

1. Electrical wavelength (feet) = $984 \frac{V}{f} = \frac{300\ V}{f}$ = (meters)
   where V = velocity factor
   f = frequency in MHz
2. This stub is a quarter-wavelength on 15 meters.
3. This stub is a quarter-wavelength on 20 meters.
4. Compensating stub.
5. TVI prevention.

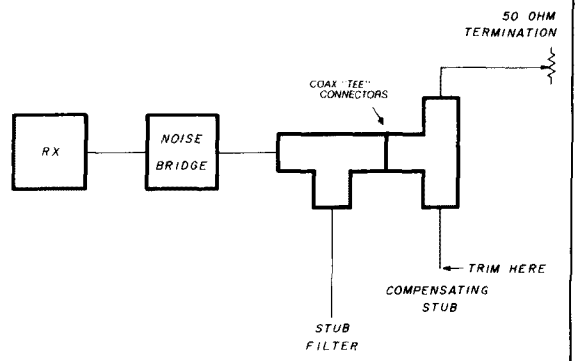


fig. 2. If you need a compensating stub, tune the receiver to the operating band (for example, 40 meters, after the stub has been tuned as a 15-meter quarter-wave), add the compensating stub, and then trim the compensating stub as before. Some compensating stubs are open-circuited, some are short-circuited, so be careful which one is used.

capacitive reactances cancel.) In general, what you do to tune a filter is adjust the half- or quarter-wave stub by itself and then add and trim the compensating stub.

trimming to length

Figure 1 shows the test setup used to trim the stub filters. To make a filter, cut the stubs to the lengths listed in **table 1** and add 10 percent to allow for variation in the velocity factor of the coax. Put a connector on one end and open or short-circuit the free end as required by **table 1.**

In adjusting the stubs, you have two cases to deal with: open-circuited half-wave stubs and short-circuited quarter-wave stubs. Trim the open-circuited stub, at the center frequency of the band for which it is a half-wavelength, until the noise bridge reads 50-ohms resistance with zero reactance.

For case of the shorted stub, trim it in the same way on the center frequency of the band on which it is a quarter wave. This trimming may simply involve moving a shorting pin down the cable until you reach the 50-ohm point. Remember that what you're really doing in this exercise is establishing an integral number of quarter-wavelengths of coax — by short- or open-circuiting a stub, you can treat it as a half- or quarter-wave on a convenient frequency — to arrive at the right length.

When you add a compensating stub (see **fig. 2**), trim its length to bring the noise bridge reading to 50-ohms resistance. This trimming adjusts the compensating stub's reactance to cancel out the filter stub's reactance.

velocity factor varies

Electromagnetic energy travels more slowly in a dielectric than it does in free space. The velocity factor is a constant that scales wavelength to reflect the media in which the energy is traveling. Because air is essentially free space, open-wire line with its air-dielectric has a velocity factor of 0.97.

Foam-dielectric coax is mostly air, so its velocity factor is about 0.8. In contrast, solid-dielectric coax, containing practically no air, has a velocity factor of 0.66. One word of caution: although foam-dielectric coax has less loss than solid-dielectric coax, its velocity factor varies with the amount of air in the foam. Consequently, the velocity factor can vary from roll to roll or from section to section of coax within the same roll. Don't assume the velocity factor for any two pieces of foam coax is the same. Always allow extra length in a stub before trimming it to size.

bibliography

Jordan, Edward C. and Balmain, Keith G., Electromagnetic Waves and Radiating Systems, 2nd Edition, Prentice-Hall Inc., Englewood Cliffs, New Jersey, 1968, Chapter 7.

The Radio Amateur's Handbook, 49th Edition, American Radio Relay League, Newington, Connecticut, 1972, Table 20-1, page 569.

ham radio

L-band ground wave propagation program

Update your VHF/UHF propagation program to include three additional ham bands

The increasing popularity of VHF/UHF communications has prompted us to expand our computer programs, described previously in *ham radio*, to include three additional ham bands.[1] If you have a working copy of the VHF propagation prediction program listed in our earlier article, you can easily update it to include the 1215-1300, 2300-2450, and 3300-3500 MHz ham bands by using the modifications described below.

L-band propagation data

The data for propagation predictions we used is taken from the ESSA Technical Report.[2] The curves from which we derived the L-band data are for 1.6 GHz (1600 MHz), but the program scales any input frequency from 1.0 to 3.5 GHz by a 20 log (f/f_{REF}) factor to accomodate all three ham bands within this frequency range. Although the original program included ground-to-air and air-to-air data, we have included only ground-to-ground data here, since most of you will not be working these bands aeronautical mobile.

entering the program

It's not necessary to type the entire new program into your computer.

If you have a Commodore 64 or 128PC, you can easily add the new data to your old program. First load your old program, then type the new lines listed in **fig. 1**, hitting <RETURN> at the end of each new line typed. The BASIC editor will automatically replace the old line numbers in the program with the new ones, and the new data lines (**4000-4610**) will be automatically added to the end of the program. Now save the new program to either tape or disk (using a new program name such as "Propagation V2.0"), and you have a propagation program covering 100 MHz through 3.5 GHz.

Changes to the TI-99/4A program are similar, but because of the limited RAM in this computer, you'll need a separate program for each band (see previous article). Load your VHF program and then do the following:

- Delete the VHF data (**lines 2005-2610**)
- In **lines 14, 38,** and **430**, change "VHF" to "L-BAND"
- In **lines 38** and **426**, change "100" to "1000"
- In **lines 40** and **428**, change "175" to "3500"
- In **line 70**, change "GOSUB 2005" to "GOSUB 4005"
- In **line 258**, change "100-175" to "1000-3500"
- In **line 424**, change "125" to "1600"

Now add the L-band data statements (lines **4005-4610**) from the Commodore program listing in **fig. 1** and save your new program.

If you've entered our previous program into another brand of computer, make the appropriate changes for your computer.

using the programs

Use of the new version is no different than use of the previous VHF/UHF programs, except that only ground-to-ground data has been included.[1] (When the program is running and you're asked to select your antenna height from the menu, select only heights of 25, 50, 100, or 500 feet.)

By Lynn A. Gerig, WA9GFR, R.R. #1, Monroeville, Indiana 46773, and **Joseph R. Hennel**, 4316 Winston Drive, Ft. Wayne, Indiana 46806

```
10 PRINTCL$:PRINT"   VHF/UHF/L-BAND PROPAGATION PROGRAM
14 PRINT:PRINT"            V2.0 C 1985 BY
50 PRINTCL$:PRINT"THIS PROGRAM CALCULATES EXPECTED RANGES
52 PRINT"FOR VHF (100-175 MHZ) UHF (225-400 MHZ)
54 PRINT"AND L-BAND (1-3.5 GHZ).  APPROXIMATE
100 PRINTCL$"PRESS (V) FOR VHF, (U) FOR UHF,          OR (L) FOR L-BAND":PRIMT
108 IFF$="L"THENPRINT"ENTERING L-BAND DATA":GOSUB4000:GOTO200
110 GOTO102
300 PRINTCL$;:IFF$()"L"THEN PRINTF$"HF PROPAGATION:  FREQ ="F"MHZ
301 IFF$="L"THEN PRINT"L-BAND PROPAGATION:  FREQ ="F"MHZ
4000 F1=1600:FL=1000:FU=3500:REM L-BAND DATA
4005 H$( 1, 1)="140010175024190050200088224208
4010 H$( 1, 2)="140012175028190057200092225212
4015 H$( 1, 3)="140014175032190060200098225218
4020 H$( 1, 4)="140030175045190070200106225225
4080 H$( 2, 2)="130009175030190060200095230240
4085 H$( 2, 3)="130010175035190065200100230245
4090 H$( 2, 4)="135027175052190078200110230258
4150 H$( 3, 3)="130015175040190065200105230250
4155 H$( 3, 4)="135035175055190078200120230260
4215 H$( 4, 4)="135040136050175067195100235300
4610 RETURN
```

fig. 1. BASIC LIST of Commodore 64 program changes.

If you don't want to make these changes yourself, or if you want a verified copy of the entire program (including the ground-to-air and air-to-air data), send a blank tape or formatted disk (tape only for the TI version) with a stamped, self-addressed return mailer and a check or money order for $5.00. If you want a full program listing, send an SASE (with 2 first-class units of postage) and $1.00. Send all Commodore requests to Lynn A. Gerig, and all TI requests to Joseph R. Hennel, at the addresses listed at the beginning of this article.

references

1. Lynn A. Gerig, WA9GFR, and Joseph R. Hennel, "Trade off Power for Antenna Gain at VHF?," *ham radio*, July, 1985, page 32.
2. *ESSA Technical Report ERL 111-ITS 79, Transmission Loss Atlas for Select Service Bands from 0.125 to 15.5 GHz*, Institute for Telecommunication Sciences, Boulder, Colorado, May, 1969. (Available for $1.25 from Superintendent of Documents, United States Government Printing Office, Washington, D.C. 20402.)

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Tell 'em you saw it in HAM RADIO!

designing Yagis
with the Commodore 64

Add screen enhancements to original WA3EKL program

Last summer, *ham radio* published WA3EKL's article, "Designing Yagis with the Commodore 64" (June, 1985, page 59). While running this technically excellent program, it occurred to me that certain features could be added to the program to enhance the quality of the visual display.

As is, the program may be confusing — for example, when the primary input of the frequency is suggested, input appears at the lower boundary of the screen and breaks into the next line, while the upper part of the screen continues to display the unnecessary data.

table 1. Summary of Commodore graphics commands. When you see bracketed commands [ ] in the instructions, press the keys as indicated. *Do not type in brackets*!

[CLR]	Press **SHIFT & CLR HOME**
[5 SPC]	Press **SPACE BAR** five times, or as many times as indicated by the number
[RVS ON]	Press **CTRL & 9**
[RVS OFF]	Press **CTRL & 0**
[4 DOWN]	Press **CRSR DOWN** four times, or as many times as indicated by the number shown.

Figure 1 lists WA3EKL's program, with the general sections rewritten to provide highlighted information and clear data input screens. **Table 1** provides a summary of Commodore graphics commands, to simplify recall of specific symbols.

Just follow the instructions step by step and you'll enjoy an "easy on the eyes" computer program that will allow you to ignore the program and concentrate instead on the dimensions of your next antenna project.

instructions

Load your previously typed WA3EKL program from disk or cassette, but *do not run* the program. Then type in the enhancement program.

Note that the numbered lines of the enhancement program will automatically replace the existing lines of the WA3EKL program memory. Some new lines have also been added.

Spacing is important. Type each line *exactly* as shown. You'll notice that some words are run together; this is intentional — doing so places them in the proper location on the screen.

You'll have to use the symbol ? instead of typing **PRINT** for some lines to get the computer to accept the long line. (See your copy of *The Commodore User's Guide.)*

Line 80 changes the screen/border/letter colors. If you like the standard blue, omit all the **POKE** statements, and type in only **PRINT "[CLR]"**.

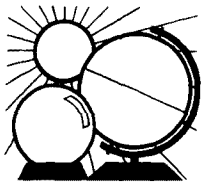
When you've finished typing, save the enhanced program under a new name of your choice.

By Fred A. Sontag, NØCAO, Lake Farm, Rt. 1, Box 47, Tebbetts, Missouri 65080

```
80 POKE53280,13:POKE53281,11:POKE646,7:PRINT"[CLR]"
100 PRINT"******* MONOBAND ANTENNA DESIGN ********":PRINT
106 IFC$="N"THEN311
108 PRINT:PRINT"[CLR]"
110 PRINT"THIS PROGRAM WAS WRITTEN USING THE
111 PRINT"FORMULAS PUBLISHED IN 1980 HAM RADIO
112 PRINT"MAGAZINE AND WRITTEN BY W2PV WHO PROVED MATHEMATICALLY AND EXPERIMENTA
LLY
114 PRINT"THAT ANTENNA ELEMENT RESONATE FREQUENCY LENGTH IS NO SIMPLE CALCULATIO
N"
116 PRINT"ESPECIALLY WHEN TAPERED ELEMENTS ARE[4 SPC]INVOLVED."
130 PRINT"****** MECHANICAL RECOMMENDATIONS ******":PRINT
140 PRINT"20 METER BOOM SHOULD BE 3 OR 4 INCH"
141 PRINT"DIAMETER ALUMINUM IRRIGATION PIPE"
150 PRINT"DEPENDING ON LENGTH."
160 PRINT"ELEMENTS SHOULD BE TAPERED SEGMENTS FROM1 1/4 INCH TO 1/2 INCH"
184 PRINT"[DOWN]"
185 PRINT:PRINT"TO CONTINUE INSTRUCTION PRESS [RVS ON]RETURN[RVS OFF]";:INPUTI$
186 PRINT"[CLR]"
190 PRINT:PRINT"10-15 METER BOOMS SHOULD BE MADE OF[5 SPC]2 INCH ALUMINUM IRRIGA
TION PIPE.
200 PRINT"ELEMENTS SHOULD BE TAPERED FROM 7/8 INCHTO 5/8 INCH DIAMETERS OF"
215 PRINT"FOR MAXIMUM FRONT TO BACK RATIO USE 1/4 WAVELENGTH BOOM ON A 3 ELEMENT
"
220 PRINT"BEAM AND 3/4 WAVELENGTH BOOM ON A 4,5[3 SPC]OR 6 ELEMENT BEAM."
225 PRINT"EQUALLY SPACE ALL ELEMENTS ALONG THE[4 SPC]BOOM REGARDLESS OF LENGTH."
230 PRINT"FOR A GOOD 2 ELEMENT BEAM USE A BOOM[4 SPC]LENGTH OF .150 WAVELENGTH."
240 PRINT"FOR A GOOD 3 ELEMENT BEAM USE A BOOM[4 SPC]LENGTH OF .300 WAVELENGTH."
244 PRINT"[4 DOWN]"
245 PRINT:PRINT"TO CONTINUE INSTRUCTION PRESS [RVS ON]RETURN[RVS OFF]";:INPUTI$:
PRINT"[CLR]"
250 PRINT:PRINT"IMPORTANT ALL TOTAL LENGTH DIMENSIONS[3 SPC]WILL BE ELEMENT HALF
LENGTH,"
255 PRINT"MEANING THE LENGTH OF THE ELEMENT[7 SPC]MEASURED FROM THE BOOM OUT TO
THE"
260 PRINT"TIP OF THE ELEMENT.ENTER ELEMENT SEGMENTLENGTHS AND DIAMETERS JUST"
270 PRINT"(WHICH CROSSES OVER THE BOOM)TO THE[5 SPC]SMALLEST DIAMETER SEGMENT AT
 THE"
275 PRINT"ELEMENT TIP. THE OBJECT IS TO ADJUST THEORIGINAL SEGMENT LENGTHS SO"
280 PRINT"THAT THE TOTAL EQUIVALENT LENGTH IS[5 SPC]EQUAL TO OR VERY CLOSE TO TH
E"
285 PRINT"NORMALIZED LENGTH FOR THE PARTICULAR[4 SPC]ELEMENT YOU ARE WORKING ON.
"
289 PRINT"[3 DOWN]"
290 PRINT:PRINT"TO CONTINUE INSTRUCTION PRESS [RVS ON]RETURN[RVS OFF]";:INPUTI$
291 PRINT"[CLR]"
295 PRINT"WHEN THE COMPUTER ASKS FOR INFORMATION[2 SPC][DOWN][RVS ON]-ENTER IN T
HE INFORMATION-[RVS OFF]"
296 PRINT
300 PRINT"THEN PRESS [RVS ON]RETURN[RVS OFF] - NOTHING WILL HAPPEN [DOWN]UNTIL[2
 SPC]YOU PRESS [RVS ON]RETURN[RVS OFF]"
301 PRINT
305 PRINT"IF NOTHING HAPPENS IMMEDIATELY,WAIT --- [DOWN][RVS ON] THE COMPUTER MA
Y BE THINKING![RVS OFF]"
309 PRINT"[9 DOWN]"
310 PRINT:PRINT"TO START CALCULATIONS PRESS [RVS ON]RETURN[RVS OFF]";:INPUTI$
311 PRINT"[CLR]"
450 PRINT:PRINT"ENTER CENTRAL DESIGN FREQUENCY IN MHZ";:PRINT"[DOWN]":INPUTCDF
470 PRINT"ENTER NUMBER OF ANTENNA ELEMENTS 2,3,4,5 OR 6";:PRINT"[DOWN]":INPUTP
490 PRINT"ENTER BOOM DIAMETER";:PRINT"[DOWN]":INPUTBD
500 PRINT"ENTER NUMBER OF ANTENNA SEGMENTS";:PRINT"[DOWN]":INPUTN
550 PRINT:PRINT"ENTER (IN INCHES) LENGTH OF SEGMENT";:PRINT"[DOWN]":INPUTA
570 PRINT"ENTER (IN INCHES) DIAMETER OF SEGMENT";:PRINT"[DOWN]":INPUTB
970 PRINT:PRINTS"INCHES=TOTAL EQUIV.LENGTH
972 PRINT:PRINT"HARDCOPY (Y/N)";:PRINT"[DOWN]"
980 PRINT"WANT TO CHANGE A SEGMENT LENGTH(Y/N)";:PRINT"[DOWN]":INPUTB$
1001 PRINT"[DOWN]"
1130 PRINT"ENTER NEW SEGMENT LENGTH (IN INCHES)";:PRINT"[DOWN]":INPUTU
3030 PRINTRS;"INCHES=NORMALIZED REFLECTOR[20 SPC]1/2 LENGTH"
3040 PRINTDE; "INCHES=NORMALIZED DRIVEN EL.[19 SPC]1/2 LENGTH"
3050 PRINTDR;"INCHES=NORMALIZED DIRECTORS[20 SPC]1/2 LENGTH"
4000 PRINT:PRINT"[RVS ON]PLEASE WAIT ..... I'M THINKING[RVS OFF]"

READY.
```

fig. 1. Yagi design screen enhancement program listing.



DX FORECASTER

Garth Stonehocker, KØRYW

MUF patterns and trends

For years computer programs have been used to provide the engineering (design) and operating maximum usable frequencies (MUF) necessary for a radio link between communications stations. These calculations define the equipment design and operation based on the minimum and maximum of sunspot cycle (years), season (months), and night and day (hours) for the transmission distance involved. The calculated hourly MUF is statistically based on the middle value (median) available during the month. There are variations from this median value throughout the month. The 10 and 90 decile values of the distribution are 15 percent under and over the median value for 27 days (90 percent) of a 30-day month. The 90 percent reliability figure is achieved by using a frequency 15 percent below the median value. Radio Amateurs can use the 15 percent *above* MUF median value, which should be reliable 10 percent of the time or three days a month.

A study has been done on the ratios (10 and 90 percentiles) to the median MUF. The largest variations in these distributions are diurnal and seasonal. **Table 1** summarizes some of the results in terms of percent below (−%) and above (+%) for day and night deviations for three seasons (Winter, Equinox, and Summer) at propagation control point latitudes.

Interestingly, the daily MUF or foF2 making up the distribution in a monthly median is precisely the daily or next-day value forecasters are trying to determine. Some patterns and trends are worth noting. Nighttime variations from the median MUF are greater than daytime variations. The midlatitudes show lower variations and these variations rise as one approaches the equator because of the geomagnetic equator. The variations increase toward high latitudes (because of the geomagnetic pole-auroral oval) during the night. The equatorial effect is more noticeable in winter and equinox than summer. This helps explain and confirm observed transequatorial DX openings to the southern hemisphere. However, in the daytime when ion production is more stable the MUF dispersion decreases from high latitudes (70 degrees) all the way to the low latitudes (10 degrees).

table 1: Variations in the 10 percent, 90 percent ratio from median MUF values.

latitude	season	ratio interval −% day	+%	−% night	+%
70	Winter	27	29	26	34
	Equinox	25	27	30	35
	Summer	19	18	22	20
50	Winter	16	13	19	23
	Equinox	24	21	24	21
	Summer	22	19	19	17
30	Winter	13	15	22	27
	Equinox	17	14	21	22
	Summer	20	23	21	20
20	Winter	16	19	25	33
	Equinox	14	14	24	31
	Summer	19	19	24	26
10	Winter	12	15	23	27
	Equinox	13	15	24	25
	Summer	16	21	24	27

At the high latitudes the increased MUF dispersion reflects the influence of solar wind particles coming into the ionosphere at the auroral oval around the geomagnetic poles. This is the region where short time variations occur (QSB), and low signal strengths associated with geomagnetic storms are received. Propagation on east-west paths to Europe and Japan are those mostly affected. However, this ionospheric movement also causes unusual propagation openings to occur. This high-latitude MUF increased distribution dispersion should be sunspot number-sensitive and next month's column will look at that.

last-minute forecast

The higher frequency bands (10-30 meters) are expected to be very good during the first and last weeks of the month. Don't look for too much in this winter season because the 27-day solar cycle isn't a big increase even though the ionosphere has increased sensitivity, as discussed last month. Little enhancement from geomagnetic field disturbances of transequatorial propagation (openings) is expected, since January is one of the most geomagnetically stable months of the year. The lower frequency bands will be best during the middle of the month. Low noise and signal absorption work together to make excellent signals, even out of the weak DX. However, abnormally weak signals (the winter anomaly) might be received for a few days this month on high midlatitude east-west and north directed paths.

Lunar perigee occurs on the 8th, with a full moon on the 26th this month. An intense but short-duration meteor shower, the Quadrantids, will occur between January 2nd and 4th and last a few hours.

band-by band summary

Ten, twelve, fifteen, and twenty meters will be open from morning to early evening almost every day, and to most areas of the world. The openings on the higher of these bands will be shorter and will occur closer to local noon. Transequatorial propagation on these bands will more likely occur toward evening during conditions of highest solar flux and a disturbed geomagnetic field.

WESTERN USA

GMT	PST	N	NE	E	SE	S	SW	W	NW
0000	4:00	40	40	20	10	10	10	10	20
0100	5:00	30	40	20	12	10	10	10	20
0200	6:00	30	40	20	20	12	10	10	20
0300	7:00	30	40	20	20	15	12	12	20
0400	8:00	30	40	20	20	15	20	20	30
0500	9:00	40	40	20	20	20	20	20	30
0000	10:00	40	*40*	*30*	20	20	20	20	30
0700	11:00	40	*40*	*30*	20	20	20	20	30
0000	12:00	40	*40*	*30*	*20*	*20*	*20*	20	40
0800	1:00	40	*40*	*30*	*30*	*20*	*20*	*30*	40
1000	2:00	40	*40*	*30*	*30*	*30*	*30*	*30*	40
1100	3:00	*40*	*40*	*30*	*30*	*30*	*30*	*30*	40
1200	4:00	*40*	*40*	*30*	*30*	*30*	*30*	*30*	*40*
1300	5:00	*40*	*40*	20	*20*	*30*	*30*	*30*	*40*
1400	6:00	*40*	30	12	12	15	*30*	*30*	*40*
1500	7:00	*40*	20	12	12	15	*20*	*30*	40
1600	8:00	40	20	10	10	15	20	*20*	*40*
1700	9:00	40	20	10	10	12	20	*20*	*80**
1000	10:00	40	30	10	10	12	15	20	*80*
1900	11:00	*40*	30	10	10	12	12	20	*80*
2000	12:00	*80*	40	15	10	12	12	15	30
2100	1:00	*80*	40	15	10	12	10	15	20
2200	2:00	*80*	40	20	10	12	10	12	20
2300	3:00	*80*	40	20	10	10	10	12*	20

MID USA

MST	N	NE	E	SE	S	SW	W	NW
5:00	40	40	20	12	10	10	15	20
6:00	40	40	20	15	15	10	12	30
7:00	40	40	20	20	15	12	20	30
8:00	30	40	20	20	20	20	20	30
8:00	30	40	20	20	20	20	20	40
10:00	30	*40*	20	20	20	20	20	40
11:00	40	*40*	*30*	20	20	20	20	40
12:00	40	*40*	*30*	*20*	*20*	20	20	40
1:00	40	*40*	*30*	*30*	*20*	*20*	*30*	40
2:00	*40*	*40*	*30*	*30*	*30*	*30*	*30*	40
3:00	*40*	*40*	*30*	*30*	*30*	*30*	*30*	40
4:00	*40*	*40*	*30*	*30*	*30*	*30*	*30*	*40*
5:00	*40*	30	20	*30*	*30*	*30*	*30*	*40*
6:00	40	20	15	20	*30*	*30*	*30*	*40*
7:00	40	20	12	12	*20*	*30*	*30*	*40*
8:00	40	20	10	12	20	*20*	*20*	*40*
9:00	40	20	10	10	20	*20*	*20*	*40*
10:00	*40*	20	10	10	20	20	20	*80*
11:00	*40*	30	10	10	12	15	20	*80*
12:00	*40*	30	10	10	12	12	20	*80*
1:00	*80*	40	15	10	12	12	15	30
2:00	*80*	40	15	10	12	10	12	20
3:00	*80*	40	20	10	12	10	12	20
4:00	*80*	40	20	10	12	10	12	20

EASTERN USA

CST	EST	N	NE	E	SE	S	SW	W	NW
6:00	7:00	40	40	20	15	15	20	12	30
7:00	8:00	40	40	20	20	20	20	20	30
8:00	9:00	40	40	20	20	20	20	20	40*
9:00	10:00	40	40	20	20	20	20	20	40
10:00	11:00	40	*40*	20	20	20	20	20	40
11:00	12:00	40	*40*	20	20	*20*	20	20	40
12:00	1:00	40	*40*	*30*	*20*	*20*	20	20	40
1:00	2:00	40	*40*	*30*	*20*	*20*	*30*	*30*	40
2:00	3:00	*40*	*40*	*30*	*20*	*20*	*30*	*30*	40
3:00	4:00	*40*	*40*	*30*	*30*	*30*	*30*	*30*	40
4:00	5:00	*40*	30	*15*	*30*	*30*	*30*	*30*	40
5:00	6:00	40	20	12	20	*30*	*30*	*30*	*40*
6:00	7:00	30	20	10	15	*20*	*30*	*30*	*40*
7:00	8:00	40*	20	10	12	20	*20*	20	*40*
8:00	9:00	40	20	10	12	15	*20*	20	*40*
9:00	10:00	40	20	10	10	15	*20*	20	*40*
10:00	11:00	*40*	20	10	10	12	20	20	*40*
11:00	12:00	*40*	20	10	10	12	20	20	*40*
12:00	1:00	*40*	20	10	10	12	15	20	*80*
1:00	2:00	*80*	30	10	10	12	15	20	*80*
2:00	3:00	*80*	40	15	10	12	12	15	*80*
3:00	4:00	*80*	40	15	10	12	10	12	20
4:00	5:00	*80*	40	20	10	12	10	12	20
5:00	6:00	*80*	40	20	12	12	15	12	20

JANUARY	ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN
WESTERN USA								
MID USA	ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN
EASTERN USA	ASIA FAR EAST	EUROPE	S. AFRICA	CARIBBEAN S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.

*Look at next higher band for possible openings.

ENGINEERING PROGRAMS FOR THE IBM-PC

RF-CAD ELECTRONICS DESIGN PROGRAM — Version 3.51 NEW!

by Joe Reisert, W1JR and Gary Field, WA1GRC

For IBM PC and compatible computers

This software package has been written by electronic engineers and contains nearly 40 tested and proven programs that will help the Radio Amateur or engineer design many common types of radio circuitry. Emphasis has been placed upon ease of use. Wherever possible, menus of choices with examples are displayed. Should the user be computer literate, the programs are not copy protected so they can be *modified to meet your specific requirements*, (full documentation is also provided.) Programs include: Fillers, LC, active- LP, HP, BP; Inductor design, toroid, solenoid, straight wire; Matching networks; Crystal oscillators; Microstrip; Transmission lines; Antennas, Yagi-Uda, helix, dish, horn, element scaling; Pi and T attenuators. Also included: Radio Path calculations; FM modulation analysis, Miscellaneous conversions; Geostationary satellite pointing; Moon tracking aids; Receiver noise figure calculations and Spurious receiver response prediction. Requires IBM-PC with at least one floppy drive and 128k of RAM.
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RF NOTES — IBM-PC

by John Simmons, W6MDI

Here's an easy way to get answers for often asked electronics questions. Both volumes contain programs written by RF consulting engineers that answer a number of very important questions often asked by Hams. Monochrome and color versions available. Written in Basic A and fully menu driven. Graphics card and 128K memory required.

Volume 1

Contains: *dB conversions*, to convert voltage, current or power levels to dB; *dBm conversions*, converts voltages or power levels to dBm and dBm to voltage or power; *VSWR calculations*, calculates VSWR and return loss when both reflected and incident powers are known; *Filter design*, 14 different filter configurations including schematics (6 low pass, 4 high pass, 2 band pass and 2 band elimination circuits; *Basic Microstrip and strip line design*, *Resonant Circuits*, design parallel and series resonant circuits, pi, capacitive and inductive impedance divider circuits.

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Volume II

This program covers: *Attenuator pads*, calculates constants for eleven different pad configurations (all with circuit diagrams.) Inductors, Inductance in a single length of wire, single layer coils, both close and wide space wound and Torroidal coil design that gives automatic selection of wire size and torroidal form. Capacitors, calculates self resonant frequencies, determines optimum bypass values and decoupling applications. Impedance Matching Networks, including, L, pi, T and series L configurations.

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Thirty and forty meters will be useful almost 24 hours a day. Daytime conditions will resemble those on 20 meters. Skip distances and signal strength may decrease during midday on days that coincide with the higher solar flux values. Nightime DX will be good except after days of high MUF conditions and during geomagnetic disturbances. Look for DX from unusual places on east, north, and west paths during this time. The usable distance is expected to be somewhat less than 20 in daytime and greater than on 80 at night.

Eighty and one-sixty meters will exhibit short-skip propagation during daylight hours and lengthen for DX at dusk. These bands follow the darkness regions opening to the east just before your sunset, swinging more to the south near midnight, and ending up in the Pacific areas during the hour or so before dawn on the path of your interest. The 160 meter band opens later and ends earlier than 80.

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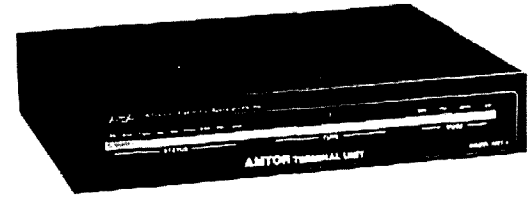
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118

NEW products

Complete Novice

"The Complete Novice" is a new package from Gordon West that contains everything the beginner needs to pass the Novice test.

Packaged in a large transparent vinyl hangbag, "The Complete Novice" contains the following study materials: four stereo code learning tapes, two stereo Novice class theory cassettes, a copy of ARRL's Novice theory book, the ARRL FCC Rule Book, a brass telegraph key, a solid-state code practice oscillator, including battery and hook-up wires, instructions for proper code sending, a list of the 200 FCC Novice test questions, a copy of FCC Form 610 plus VEC Examiner lists, a confidential written exam and ten 5-WPM tests for the examiner, and complete test instructions to the Volunteer Examiner.

The Complete Novice is available from Ham Radio's Bookstore or from Radio School, Inc., 2414 College Drive, Costa Mesa, California 92626. The price is $49.95 plus $5.00 for shipping and handling.

Circle #312 on Reader Service Card.

newest accessories for Kenwood HT

An extra-life battery pack and an AC operated quick-charger for the Kenwood TH-Series pocket transceivers are now available.

The PB-21H is an extra-life 500 mAH NiCad battery pack measuring just half an inch more than the standard PB-21 battery pack, and weighs only 6.5 ounces. (The standard PB-21 NiCad pack is rated at 180 mAH.)

The new BC-6 is an AC-operated, two-pack quick-charger that doubles as a DC power source for the TH-Series radios. The BC-6 can fully charge either the standard PB-21 or the new extra-life PB-21H in just one hour. The two-pack quick-charger comes complete with an adapter cable so you can operate your HT while the battery packs are charging.

For further information, contact Trio Kenwood Communications, 1111 West Walnut Street, Compton, California 90220.

MFJ Enterprises. Inc.

The new MFJ-818 automatic digital SWR/Wattmeter from MFJ Enterprises, Inc., is unique in several ways. First, it's digital, with easy-to-read, 1/2 inch bright-orange digits on the SWR display. But more importantly, MFJ-818 is automatic. It eliminates three steps in reading SWR: switching to set, setting the meter needle for full-scale deflection, and switching back to SWR 1:1 to 9.9:1 directly and instantaneously. There's no need to adjust the SWR set knob!

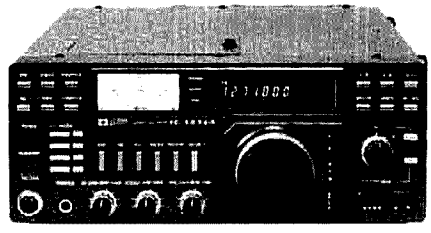
The MFJ-818 reads up to 200 watts RF output on its LED bar graph disply. The 12-bar LED display indicates the "On Air" power level instantly and correctly to read instantaneous peak power. The unit features a tri-color indicator that lights up to show the antenna matching condition: green for good, yellow for "not very good," and red for a mismatched condition.

Despite all the features packed into this model, it measures only 5-1/2 x 4-1/4 x 1 inches and retails for only $89.95 each. Like all MFJ products, the MFJ-818 carries a one-year unconditional warranty. In addition, if you order directly from MFJ, you get a 30-day money back guarantee: if you're not completely delighted, just send the SWR/wattmeter back within 30 days for a full refund (less shipping). For information, contact MFJ Enterprises, Inc. P.O.Box 494, Mississippi State, Mississippi 39762

Circle #310 on Reader Service Card.

ICOM IC-1271A 1.2 GHz transceiver

ICOM has announced the IC-1271A full-featured base station transceiver. With coverage from 1240 to 1300 MHz, the IC-1271A features 10 watts of RF output power, 32 memories, scanning and multi-mode operation including ATV (Amateur TV).



Additional features include front-end GaAs FETs for exceptional receiver sensitivity; CW/FM/upper and lower SSB; scanning (memory, program or mode scan); and 12 VDC or 117/240 VAC operation (optional). The suggested retail price is $999.00.

Options include the TV-1200 ATV interface unit, IC-EX310 voice synthesizer, UT-15S CTCSS encoder/decoder and IC-PS25 13.8 VDC internal power supply.

For further information, contact ICOM America, Inc., 2380 116th Avenue, N.E., Bellevue, Washington 98009-9029.

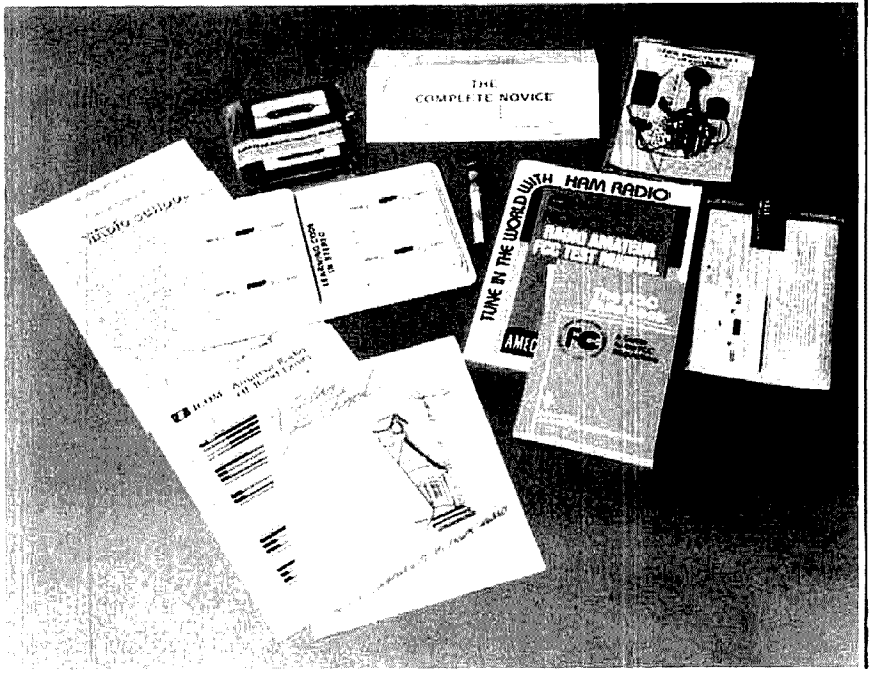
Circle #311 on Reader Service Card.

new Jensen catalog

A new 160-page catalog of hard-to-find tools, service kits, and test equipment is offered free by Jensen Tools Inc. Illustrated in full color, this catalog contains more than 1000 items, including an expanded line of circuit board equipment.

To obtain a free copy, contact Jensen Tools Inc., 7815 South 46th Street, Phoenix, Arizona 85044.

Circle #309 on Reader Service Card.



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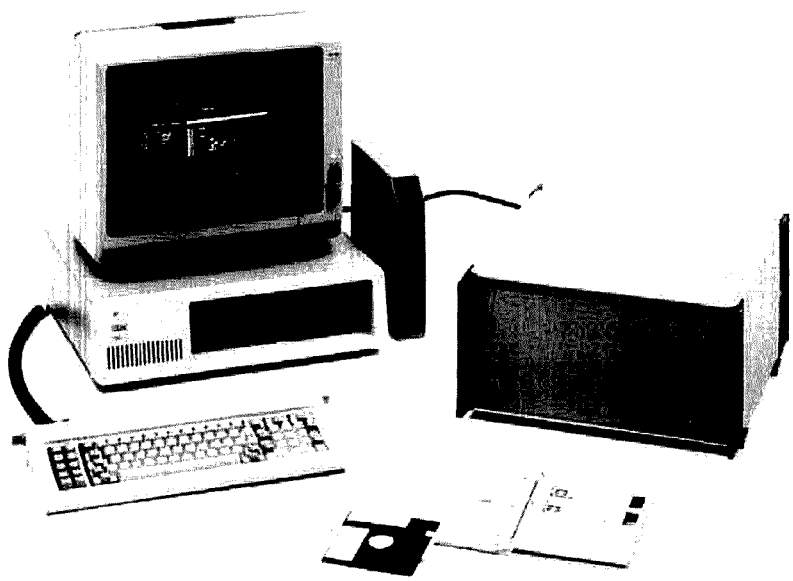
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DESCRIPTION	PRICE	DESCRIPTION	PRICE
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2E26	3.75	6HS6	2.50
3B28	6.25	6J6	2.25
4CX150A	25.00	6SH7	2.25
50C5	1.70	6TF4	22.00
5663	4.00	6W6	2.00
5749	2.00	6087	3.15
5847	6.80	6201	1.90
6AC7	2.50	6360	4.00
6AG7	2.05	6662	1.15
6AL5	2.22	6676	1.00
6AN4	2.50	12AU7	.80
6AQ6	1.45	6907	35.00
6AS5	4.50	7A8	3.75
6AU6A	2.95	7C7	3.75
6AX4GTB	5.65	7054	2.10
6BC4	3.15	7058	1.50
6BH8	2.18	7061	2.20
6BL8	1.70	7167	1.95
6BW4	2.75	7360	6.50
6C4	2.30	7716	2.25
6CB6	2.23	7868	1.95
6CH8	2.44	866A	4.75
6CM6	2.00	8102	1.75
6CX8	2.44	8348	1.25
6DJ8	1.25	8595	15.00
6DZ4	1.60	12AB5	1.60
6DT8	1.60	12AT6	1.60
6D10	1.90	12AU7	1.65
6EU7	2.10	12BE6	1.80
6EW6	5.15	12B4A	6.60
6GE5	3.00	8624	65.00
6GW8	7.00	2D21W	1.75
6J4	2.25	3B25	8.00
6L6GL	3.25	4-125	40.00
6T8	1.41	5AQ5	6.00
6V4	6.00	5651	1.75
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7094	60.00	6BD6	1.65
7271	60.00	6BS8	1.60
7701	3.75	6C10	1.75
7724	1.75	6CB5	2.15
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8077	2.00	6CL8	3.42
8167	120.00	6CS7	1.90
8505A	43.00	6C30	11.00
9002	3.25	6GHA8AG	1.65
12AQ5	1.75	6EA8	2.70
12AV7	2.04	6EA8	5.85
12BH7	6.00	6EB8	2.50
12BY7	5.10	6EV7	1.85
12X4	1.00	6FV8	2.00
2D21	1.12	6GN8	2.80
2050A	2.25	6HS8	2.50
35W4	5.00	6KE8	8.60
100TH	62.50	6SK7	3.00
5R4G	1.75	6U8	3.00
5678	4.50	6X4	1.75
5751	2.00	6B6	1.75
5963	1.90	6252	45.00
6AF4	2.20	6397	7.50
6AK5	1.55	6663	1.00
6AM4	2.24	6677	1.00
6AN5	2.40	6688	4.50
6AG7	2.52	7A6	3.75
6AS6	2.50	7D5	3.75
6AU8	2.76	7H7	3.75
6BA6	1.20	7055	2.25
6CE5	1.70	7059	1.95
6BJ6	2.70	7077	46.00
6BN6	1.80	7258	2.20
6BZ6	2.75	7377	40.00
6CA4	3.00	7717	2.25
6CE5	1.69	816	7.50
6CL6	3.95	8058	12.00
6CM8	1.80	8136	1.25
6CY5	2.45	8458	15.00
6DS4	4.80	8883	45.00
6DT6	1.70	12AL5	2.23
6DV4	3.65	12AT7	3.21
6EM5	2.15	12BA7	1.50
6EU8	1.80	12BR7	2.50
6FG5	2.50	12DW7	2.50

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- Tubes are from surplus stock.
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NEW products



new software from Fluke

John Fluke Mfg. Co., Inc. has announced a new software package that links an IBM PC, XT or AT with the Fluke 2400B Intelligent Computer Front End. ProLink PC™ Program Development Package simplifies creation of data acquisition and control applications that, once developed, are executed on the 2400B, freeing the host PC for other tasks.

The package's main benefit is the ability to turn an IBM PC to that one task. ProLink allows the PC to download its application program to the 2400B. The Fluke 2400B in turn handles all A/D conversions, linearizations, limit checking and data buffering.

The friendly programming environment provides a set of menu choices that allows the operator to create and edit 2400B measurement and control programs using menu selection. Menu choices allow programs to be downloaded to and retrieved from the 2400B. Additional menu choices allow the user to check programs for errors and execute these programs while monitoring 2400B variables and data on the PC's screen. Up to 100 channel systems can be supported by this system. The user can also select the desired IBM serial port, set the port characteristics, or exit to PC-DOS.

ProLink PC contains everything needed to get an IBM-PC and 2400B working together. Included is a floppy diskette, manual, and RS-232-C cable. ProLink PC can be ordered immediately, with a normal delivery time of six weeks. The price of the ProLink PC package is $450.

For additional information, contact John Fluke Mfg. Co., Inc., at P.O. Box C9090, Everett, Washington 98206.

Circle #308 on Reader Service Card.

new phone patch

A unique phone patch has been introduced by the Heath Company, one of the world's largest manufacturers of high technology kit products. The HD-1515 Phone Patch kit employs a unique design and special speech transmission circuits to transfer audio signals between a telephone line and two-way radio equipment.

New active circuitry uses automatic gain control to help compensate for the varying attenuation introduced by different line lengths or loop distance on the external phone circuit. The integrated active speech and transmission circuit also allows the Phone Patch to be directly connected to the phone line, thereby eliminating



conventional hybrid transformers that convert 4-wires to 2-wires.

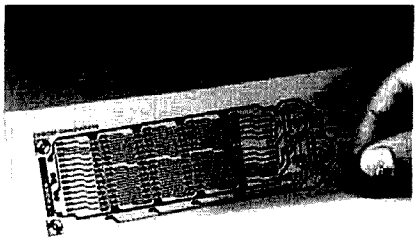
A high degree of electrical isolation from the radio equipment makes the Heathkit Phone Patch ideal for voice-operated installations. In addition, the transmission can be monitored on the phone line. An 8-pole filter in the input circuit makes the Patch compatible with all subscriber loop-frequency voice standards as prescribed by the FCC.

For information contact Heath Company, Dept. 150-575, Benton Harbor, Michigan 49022.

Circle #301 on Reader Service Card.

film simplifies making PCBs at home

Anyone who's ever made a PCB at home knows that applying the etch resist takes patience, skill, and experience.



But TEC-200™ Image Film, a special plastic film available from Meadowlake, simplifies the process to three easy steps: first, the desired circuit pattern is photocopied directly from any magazine or book, using a sheet of TEC-200 film in the paper tray of any standard "plain paper" copier that uses toner. The high heat resistance of TEC-200 and the quality of modern photocopiers result in excellent reproduction with minimal pattern distortion.

Next, the photocopied circuit pattern is transferred to a piece of copper-clad board using a hot iron. The toner deposited on the photocopy melts and forms a varnish-like, acid-proof, etch-resist coating. Just remove the film from the board, and the board is ready for etching.

For information about TEC-200 Image Film, contact The Meadowlake Corporation, 25 Blanchard Drive, Northport, New York 11768.

Circle #302 on Reader Service Card.

curlycode™

Minds Eye Publications has announced a new publication: Curlycode™, a new way to learn Morse code.

Not "just another code course" but instead a new learning experience, it lets you become familiar with all the characters in one quick reading of the manual. It shows you how to let the rhythm of the code build shapes that you see in your mind's eye. Each sound adds to the shape you already "see," so you never have to change your mind about the shape. Each shape is the character, so you can write it down instantly.

The beginner will remember most of the shapes without any study because only eight basic shapes are needed to know half of everything and nearly all the letters! Remember two additional endings for each shape and you'll know all 61 characters.

If you're already an expert, you'll find Curlycode an exciting new way to increase receiving speed. With a little practice, you may soon find the Curlycode shapes forming automatically in your mind's eye at practically any speed.

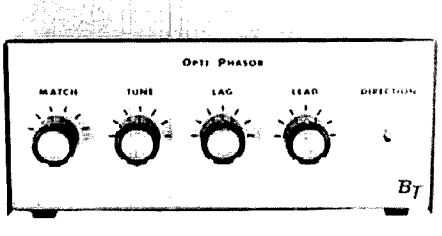
The price of the full set (including manual, wall chart, pocket card, and beginner's chart) is $11.50. The manual alone (without the separate charts) is available for $6.50. You or your club can save money by buying ten or more complete sets and deducting 20 per cent. All prices include postage and handling.

For further information, contact Minds Eye Publications, Dept. PHI, Box 1310, McLean, VA 22101.

Circle #303 on Reader Service Card.

antenna phasing unit

BaileyTech has introduced a new type of antenna product — the Opti-Phasor, an in-the-shack 40-meter phasing unit designed to drive a pair of dipoles or inverted vees to obtain a directional pattern and 4 dB gain. Either of two directions may be selected by merely throwing a switch. Variable reactance phasing allows the currents to be precisely balanced in the two dipoles so that deep nulls off the back and optimum gain can be achieved.



A front-to-back ratio of 20 dB or more is typical and the null is steerable. The five controls are "lead," "lag," "match," "tune," and the "direction" switch. Standard female UHF connectors are provided on the back of the phasor for the transmitter and two feedlines. A separate 52-ohm coax feedline is needed to drive each dipole. The Opti-Phasor will match a 52-ohm transmitter at full legal power. An SWR indicator or reflected power meter is needed to adjust the match. The user-supplied dipoles are hung parallel about 25 feet apart. Size is 7 1/2 x 3 3/8 x 6 1/8 inches; color is beige with black. The list price is $120.

For details, contact BaileyTech, 304 West S. College Street, Yellow Springs, Ohio 45387.

Circle #304 on Reader Service Card.

RF TRANSISTORS

FRESH STOCK - NOT SURPLUS
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2-30 MHz 12V (= 28V)*

P/N	Rating	Each	Match Pr
MRF406	20W	$14.50	$32.00
MRF412,/A	60W	18.00	45.00
MRF421	100W	25.00	56.00
MRF421C	110W	—	60.00
MRF422*	150W	38.00	82.00
MRF426,/A*	25W	18.00	42.00
MRF428**	150W	55.00	125.00
MRF433	12.5W	12.00	30.00
MRF435*	150W	42.00	90.00
MRF449,/A	30W	12.50	30.00
MRF450,/A	50W	14.00	31.00
MRF453,/A	80W	15.00	35.00
MRF454,/A	80W	16.00	36.00
MRF455,/A	60W	12.00	28.00
MRF458	80W	20.00	46.00
MRF460	60W	18.00	42.00
MRF464*	80W	25.00	60.00
MRF466*	40W	18.75	48.00
MRF475	12W	3.00	9.00
MRF476	3W	2.75	8.00
MRF477	40W	11.00	25.00
MRF479	15W	10.00	23.00
MRF485*	15W	6.00	15.00
MRF492	90W	18.00	40.00
SRF2072	75W	15.00	33.00
SRF3662	110W	28.00	60.00
SRF3775	75W	15.50	34.00
SRF3795	85W	16.50	37.00
CD2545	50W	23.00	52.00
SD1076	70W	17.00	40.00
SD1451	50W	15.00	36.00

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VHF/UHF TRANSISTORS

	Rating	MHz	Net Ea.	Match Pr
MRF212	10W	136-174	$16.00	—
MRF221	15W	136-174	10.00	—
MRF222	25W	136-174	14.00	—
MRF224	40W	136-174	13.50	32.00
MRF231	3.5W	66-88	10.00	—
MRF234	25W	66-88	15.00	39.00
MRF237	4W	136-174	3.00	—
MRF238	30W	136-174	12.00	—
MRF239	30W	136-174	15.00	—
MRF240	40W	136-174	18.00	—
MRF245	60W	136-174	28.00	65.00
MRF247	75W	136-174	27.00	63.00
MRF250	50W	27-174	20.00	46.00
MRF260	5W	136-174	7.00	—
MRF261	10W	136-174	9.00	—
MRF262	15W	136-174	9.00	—
MRF264	30W	136-174	13.00	—
MRF607	1.75W	136-174	3.00	—
MRF641	15W	407-512	22.00	—
MRF644	25W	407-512	24.00	54.00
MRF646	40W	407-512	26.50	59.00
MRF848	60W	407-512	33.00	69.00
2N3866*	1W	30-200	1.25	—
2N4427	1W	136-174	1.25	—
2N5591	25W	136-174	13.50	34.00
2N5642*	20W	30-200	13.75	34.50
2N5945	4W	407-512	10.00	—
2N5948	10W	407-512	12.00	—
2N6080	4W	136-174	6.25	—
2N6081	15W	136-174	7.50	—
2N6082	25W	136-174	8.90	—
2N6083	30W	136-174	9.30	24.00
2N6084	40W	136-174	11.75	28.50

TMOS FET

MRF134*	5W	2-200	$10.50	—
MRF137*	30W	2-200	22.50	—
MRF138**	30W	1.5-150	35.00	—
MRF150**	150W	1.5-150	80.00	—
MRF172*	80	2-200	65.00	—

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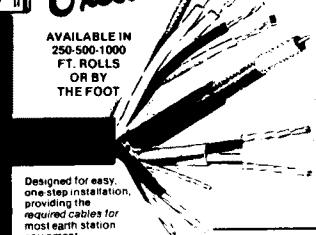


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2 @ 16 Gauge	2 @ 12 Gauge	2 @ 12 Gauge	2 @ 12 Gauge	2 @ 12 Gauge
5 @ 22 Gauge	3 @ 18 Gauge	6 @ 18 Gauge	3 @ 18 Gauge	3 @ 18 Gauge
3 @ 20 Gauge Shielded plus Tinned Copper Drain Wire	3 @ 20 Gauge Shielded plus Tinned Copper Drain Wire	3 @ 22 Gauge Shielded plus Tinned Copper Drain Wire	3 @ 20 Gauge Shielded plus Tinned Copper Drain Wire	3 @ 22 Gauge Shielded plus Tinned Copper Drain Wire
	3 @ 22 Gauge Shielded plus Tinned Copper Drain Wire		1 @ 22 Gauge Shielded plus Tinned Copper Drain Wire	
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2.54 dB/100 ft. @ 1000 MHz .....$3.25/ft.

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(FLC12-50J) 2.78 dB/100 ft. @ 1000 MHz .....$1.59/ft.

7/8 inch loss .13 dB/100 ft. @ 30 MHz
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Cable — $6.00 per 100 ft.
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RG174/U min. 50 Ω mil spec.....10¢/ft.
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UG-255 (PL-259 to BNC).....$2.95
Elbow (M359) UHF Elbow.....$1.79
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UG-88C/U BNC Male tor RG-58, Amphenol $1.25
UG 273 BNC-PL259 Amphenol.....$3.00
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beam antenna handbook
by Bill Orr, W6SAI and Stu Cowan, W2LX

Completely revised and updated, the *Beam Antenna Handbook* includes the latest state-of-the-art antenna design. Computer generated beam dimensions for the 40, 30, 20, 17, 15, 12, 10, and VHF bands are included, eliminating the need for time-consuming math calculations. Also covered are: beam height and optimum angle of radiation, how element types and hardware effect performance, effect of nearby objects on radiation patterns, feedlines, baluns, and matching systems and much more. *ham radio* VHF columnist Joe Reisert, W1JR, and noted European VHF'er DL6WU's VHF antenna designs are covered extensively as well as NBS VHF long Yagis. 286 clearly written pages — 204 easy-to-understand illustrations make this the book to buy for beam construction. The price is $9.95 softbound.

For more information, contact Ham Radio's Bookstore, Greenville, New Hampshire 03048.

Commodore companion

Jim Grubbs, K9EI (see "Micros and VHF Beacons Transmit Messages Automatically," *ham radio*, July, 1985, page 51) has authored a new book entitled *The Commodore Ham's Companion*, a complete guide to using Commodore computers in the shack. In fourteen chapters, Jim addresses a variety of topics including selecting the right Commodore computer for your station, the basics of programming for RTTY; Morse, AMTOR, and packet; and information management strategies for logging and other record keeping chores (contests, awards, etc.)

In a clear, easy-to-read format, he explains why the Commodore machines are the easiest to interface to the rest of the world, and takes a special look at "telehamming," or connecting to Amateur Radio information sources via the telephone lines. Message Storage Operations and packet are also addressed.

Jim also lists sources of software for SSTV, satellite tracking, and other specialized applications, as well as over 80 sources of hardware and software for more general Amateur Radio applications.

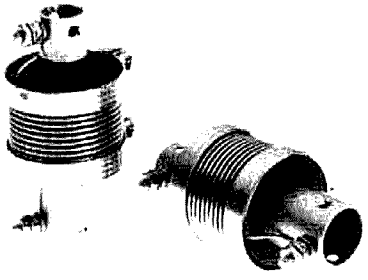
The Commodore Ham's Companion is a 160-page paperback retailing for $15.95, plus shipping and handling, ($2.50 if ordered directly from the publisher, $3.00 if ordered from *Ham Radio's Bookstore).*

For information, contact QSKY Publishing, P.O. Box 3042, Springfield, Illinois 62708.

Circle #305 on Reader Service Card.

12-meter antenna coil

KW-12 antenna coils allow trapped dipole coverage of the new 12-meter band (24.89-24.99 MHz). Resonant frequency is designed to provide a perfect half-wave dipole. The power handling capability is 1500 watts PEP maximum. The pair is installed 9 feet, 5 inches from the balun.



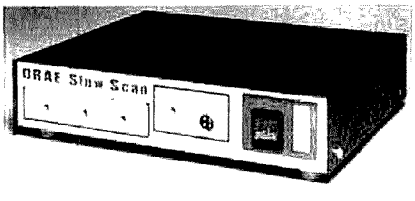
Hi-Q characteristics are obtained by optimum form factor on polystyrene. Coil dimensions are 5 1/2 inches by 1/8 inch diameter; the unit weighs 6 ounces, with tensile strength rated at 800 pounds. An acrylic lacquer waterproof coating and all-aluminum hardware help provide resistance to corrosion. A specification sheet and installation instructions are supplied.

For information, contact Microwave Filter Company, 6743 Kinne Street, East Syracuse, New York 13057.

Circle #306 on Reader Service Card.

SSTV transceiver

Davtrend has introduced two new products: the DRAE SSTV Transceiver and — for those who already have the receiver — a Transmit Module that upgrades the DRAE receiver to transceiver status.



The system specification is the standard 8.5 second SSTV format of 16 grey shades with a 128 × 128 pixel picture. Further specification details are available.

For further information, contact Davtrend Limited, Sanderson Centre, Lees Lane, Gosport, Hampshire PO12 3UL England.

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Antennas

YAESU

FT-ONE, FT-980, FT-757GX, FRG-8800, FT-726R, FRG-9600, FT-270RH, FT-2700RH

YAESU FT-270/470RH, FT-2700RH, FTC-1000 — ICOM IC2AT, IC02AT, IC04AT — Land Mobile Hi/I Motorola/Standard, Wilson, Maxon, Yaesu FTC-2203, FT-470, Icom IC-M12 Marine/M700, Tempo M-1

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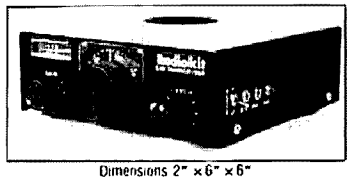
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Complete Kit

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Dimensions 2" × 6" × 6"

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Frequency	3.8-4.0 MHz
Sensitivity	0.5 μV for 10 dB S/N
Selectivity	−6 dB @ 2.4 KHz
AGC Range	+60 dB in = +3 dB out
Audio Ouput	>350 mW into 8 ohms

TRANSMITTER:

Frequency	3.8-4.0 MHz
Output	30 watts into 50 ohms
IMD	−30 dB
Harmonics	2nd — 47 dB, 3rd — 55 dB
SWR Immunity	30:1 @ all phase angles

ALC Amplified, fast response rate (quasi-processing)

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COMING EVENTS

Activities — "Places to go . . ."

CALIFORNIA: FCC exams, Novice-Extra. Sunnyvale VEC ARC. (408) 255-9000 24 hour. 73, Gordon, W6NLG, VEC.

MICHIGAN: The Southfield High School Amateur Radio Club is sponsoring their 20th annual Swap & Shop, January 26, Southfield High School, 24675 Lahser, Southfield. Doors open 6 AM for exhibitors. Public 8 AM to 3 PM. Admission $2.50. Reserved tables $20.00 (paid in advance) for two 8 ft. tables. Additional reserved tables $10.00 each. Tables also available at door. For information and reservations: Robert Younker, Southfield High School, 24675 Lahser, Southfield, MI 48034.

NEW YORK: Yonkers Amateur Radio Club's Electronics Auction, Sunday, January 26, Lemko Hall, 556 Yonkers Avenue, Yonkers. Inspection 9 AM to 10 AM. Auction starts promptly at 10 AM. Admission $3.00. Children under 8 free. Plenty of seats and parking. Unlimited free coffee all day. Club commission on successful sales only. 10% on first $100, 5% on remainder.

SOUTH BEND, INDIANA: Hamfest Swap & Shop, January 5 (first Sunday after New Year's Day) Century Center downtown on US 33 oneway North between St. Joseph Bank Building and river. Industrial history museum in same building. Carpeted half acre room. Open tables $1 per ft. Four lane highways to door from all directions. Talk in freq: 52-52, 99-39, 93-33, 69-09, 145.29.

MASSACHUSETTS: The MIT UHF Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday January 22, 1986, 7 PM, MIT Room 1-134, 77 Mass Ave, Cambridge, MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-5820/646-1641 or Craige Rodgers at 225-6616. Exam fee $4.00. Bring copy of current license, 2 forms of picture ID and completed form 610 (available from FCC in Boston. 223-6609).

WISCONSIN: The Wild Rivers Amateur Radio Club will hold their mid-winter swapfest, Sunday, February 16, 10 AM to 3 PM. Spooner Experimental Farm east of Spooner on highway 70. VE exams by advance registration only. Tables available. Talk in on 147.81/21. For information: Tom Young, KD9FC, Route 5, Box 5239, Hayward, WI 54843.

OHIO: The Mansfield Mid-Winter Hamfest/Auction, Sunday, February 16, Richland County Fairgrounds, Mansfield. Doors open to public 7 AM. Tickets $3.00 advance and $4.00 at door. Tables $5.00 advance and $6.00 at door. Talk in on 146.34/94, W8WE. For information/tickets/tables SASE to Dean Wrasse, KB8MG, 1094 Beal Road, Mansfield, Ohio 44905 or phone (419) 589-2415 after 3 PM EST.

MASSACHUSETTS: The Algonquin ARC will hold its annual Hamfest/Swapfest Electronic Flea Market, Sunday, February 16, Marlboro Junior High School, Marlboro. Doors open 8:30

AM for sellers, 10 AM for buyers. Admission $1.00. Tables $7.50 advance (by February 9) or $10.00 at the door. Food available. For information/table reservations: AARC, PO Box 258, Marlboro, MA 01752 or call (617) 393-9920.

PENNSYLVANIA: Amateur Radio Social. The Chaverim of Delaware Valley will hold their 6th annual mid-winter social in downtown Philadelphia on Sunday, January 12, at 10 AM. Traditional buffet brunch, installation of new officers, entertainment and lots of eye-balling. Everyone invited. Reservations a must. Contact Bill Soble, W3OXT (215) 676 6769 or write to 9357 Hoff Street, Philadelphia, PA 19115 for full details.

OHIO: Dayton Hamvention, April 25, 26, 27, Hara Arena and Exhibition Center. For more information write or phone Box 44, Dayton, OH 45401 (513) 433-7720. See display ad this issue.

FLORIDA: Citrus County Ham Flea Market, February 8, 9 AM to 4 PM, County Fairgrounds, US 41 south, Inverness. License exams. For information: SHARC Hamfest, PO Box 572, Lecanto, FL 32661 or call Bob Gordon, W1KUL (904) 628-5045.

OPERATING EVENTS

"Things to do . . ."

1986 CLASSIC RADIO EXCHANGE 2100 UTC January 26 to 0400 UTC January 27. Object: restore, operate and enjoy old radio equipment with like-minded hams. A classic radio is any equipment built since 1945 and at least 10 years old. Exchange name, RST, state/province/country, receiver and transmitter type and other interesting information. CW call "CQ CX", phone call "CQ Exchange". Send logs, comments, etc to Stu Stephens, K8SJ, 1407 Hollyrood Road, Sandusky, OH 44870. Include SASE for Classic Radio Newsletter printed January and September. 1986 Fall Classic Exchange is September 28-29.

1986 NEW HAMPSHIRE QSO PARTY sponsored by the NH Amateur Radio Association. 1900Z February 1 to 0700Z February 2 and 1400Z February 2 to 0200Z February 3. Exchange signal report and NH county or ARRL section/DXCC country. Stations may make contact on more than one band or mode. Logs must be submitted by March 25, 1986. Include large SASE for results. Send logs and comments to Mount Moriah Repeater Association c/o Bud Valcourt, N1BYQ, 19 Teague Drive, Salem, NH 03079.

MICHIGAN YL QSO PARTY sponsored by the TASYL's from 1800Z January 25 to 1800Z January 26, phone and CW. No crossband, net or repeater QSO's. Work stations once. Exchange call, signal report, QTH and TASYL # if working a member. Send logs to TASYL President, Verline Ferris, K1BV, 308 E. Harry, Hazel Park, Michigan 48030. Entries must be received by February 28, 1986.

CELEBRATE KANSAS' 125th Birthday Special Event Station KD0ZK will operate all General and Novice bands during January 1986. An 8-1/2 x 11 certificate is available for contact. Send QSL and large SASE to KD0ZK, Barry Horowitz, 715 West 5th Street, Junction City, Kansas 66441.

YORK RADIO CLUB will celebrate its 50th anniversary with several special programs in 1986. Starting January 1 through December 31, 1986, members of the YRC will work all HF bands at all class levels. Members will use their own call signs and add York Radio Club to all CQ's and contacts. Submit proof of contact, SASE and one dollar for a certificate. Send to York Radio Club, 861 Fairfield, Elmhurst, IL 60126. Include time, date, condx and member's call letters.

YL—OM CONTEST All licensed men and women operators throughout the world are invited to participate. OMs call "CQ YL" and YLs call "CQ OM". All bands may be used. No crossband operation, nets or repeater operation will count. Exchange station worked, QSO number, RS(T), ARRL section or country. Log entries must show time, band, date and transmitter power. Send logs to Mary Lou Brown, NM7N, Vice President, 504 Channel View Drive, Anacortes, WA 98221. Logs must be postmarked by March 10 and received no later than March 31, 1986.

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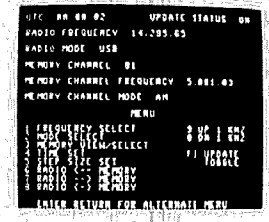
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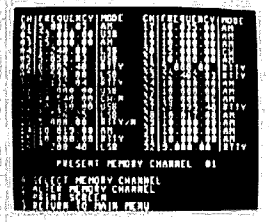
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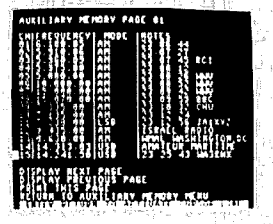
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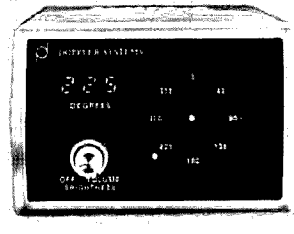
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THE GUERRI REPORT

by Ernie Guerri, W6MGI

computer graphics become more powerful

Next to word processing and accounting functions, graphics applications are rapidly becoming one of the most widely used functions of microcomputers. Unfortunately, small computers of the Apple/IBM PC class have relatively modest memory and data rate capabilities, and sophisticated graphics applications require enormous processing ability. For example, the IBM PC, using one of its most advanced graphics adapter boards, is capable of producing a picture consisting of 600 horizontal x 400 vertical picture elements (pixels) with 16 shades of gray (or 16 colors). This requires about 200,000 eight-bit bytes of memory, and can produce one full screen of picture each ⅓ second. To produce just one second of television-grade picture would require *ten* times the computer power of the PC, and to add the full 64 shades of gray that the average eye can discern would double that requirement again!

But this is only the beginning. Advanced graphics applications such as computer aided engineering (CAE) require displays with 1200 x 1000 pixels and as many as 256 different colors. This requires memories capable of storing several hundred million bytes, computers operating at 5-10 million operations/second, and CRT data rates of up to 125 MHz. Typical of the computers in this class are the DEC VAX and the Gould 32/67 — half a million dollars without the chrome and tailfins.

To produce full-color moving pictures of entertainment quality requires displays capable of 4000 x 4000 pixels with 1000 colors. The memory requirements may be in the trillions of bytes, and the computers must be capable of nearly a billion operations/second. The CRAY supercomputer is one of the few in this class — and is priced at about $10 million! LucasFilm has reportedly invested in several CRAY's for the purpose of making movies that are 100 percent computer generated. Take heart, ATVers — these goodies should be available surplus in about 50 years!

super broadband amplifiers on the horizon

We've become accustomed to the notion that "broadband" generally means that a circuit or amplifier can cover all of a given band of interest with the same characteristics that we would expect of a circuit optimized for a specific frequency. Recent advances in hybrid microwave IC's, however, are redefining the concept of "broadband." Design groups at Philips Research in England have developed hybrid ICs using MESFETs that have measured bandwidths of nearly 12 GHz with about 7 dB of gain, and a noise figure of less than 8 dB, Moreover, the unit is capable of over ¼ watt of output over the entire band! Meanwhile, researchers at Hughes aircraft report an amplifier which is flat over the range from 2-40 GHz, giving 8 dB of gain. Scientists at several other organizations are reporting similar results which would indicate great progress in the development of devices and circuit concepts. HEMT GaAs FETs promise very low noise and high gain at these frequencies, and new computer aided modeling techniques will be required to optimize the circuit layouts and structures required to yield these bandwidths. Silicon bipolar transistors still seem to have plenty of life left for these applications. The AT200 series from Avantek, used in oscillators at 20 GHz, demonstrates better stability and lower noise than GaAs designs, when followed by GaAs distributed amplifiers. Not surprisingly, the geometry and dimensions of the silicon devices is similar to the GaAs counterpart. Applications will initially be in highly sophisticated areas like electronic warfare and in the terminal ends of optical signal processors, but within a few years we should see this capability forming the basis for a whole new generation of lab and test equipment. Within a decade we could see low power transceivers that cover all the Amateur bands in a single unit.

vacuum tube ICs make debut

For the past several years investigators have contemplated the notion of using techniques that implement the microgeometries of transistors in order to make vacuum tubes with correspondingly small dimensions. The advantages would include the characteristics that make tubes attractive — high-temperature operation, radiation resistance, and substantial forgiveness of voltage surges.

Designers at Los Alamos National Laboratory have fabricated a vacuum tube "IC" that contains about 200 triodes and measures 40 mils on a side. It's expected that in about two years, devices that contain four to five times as many "tubes" and require only one-fourth as much space will be possible.

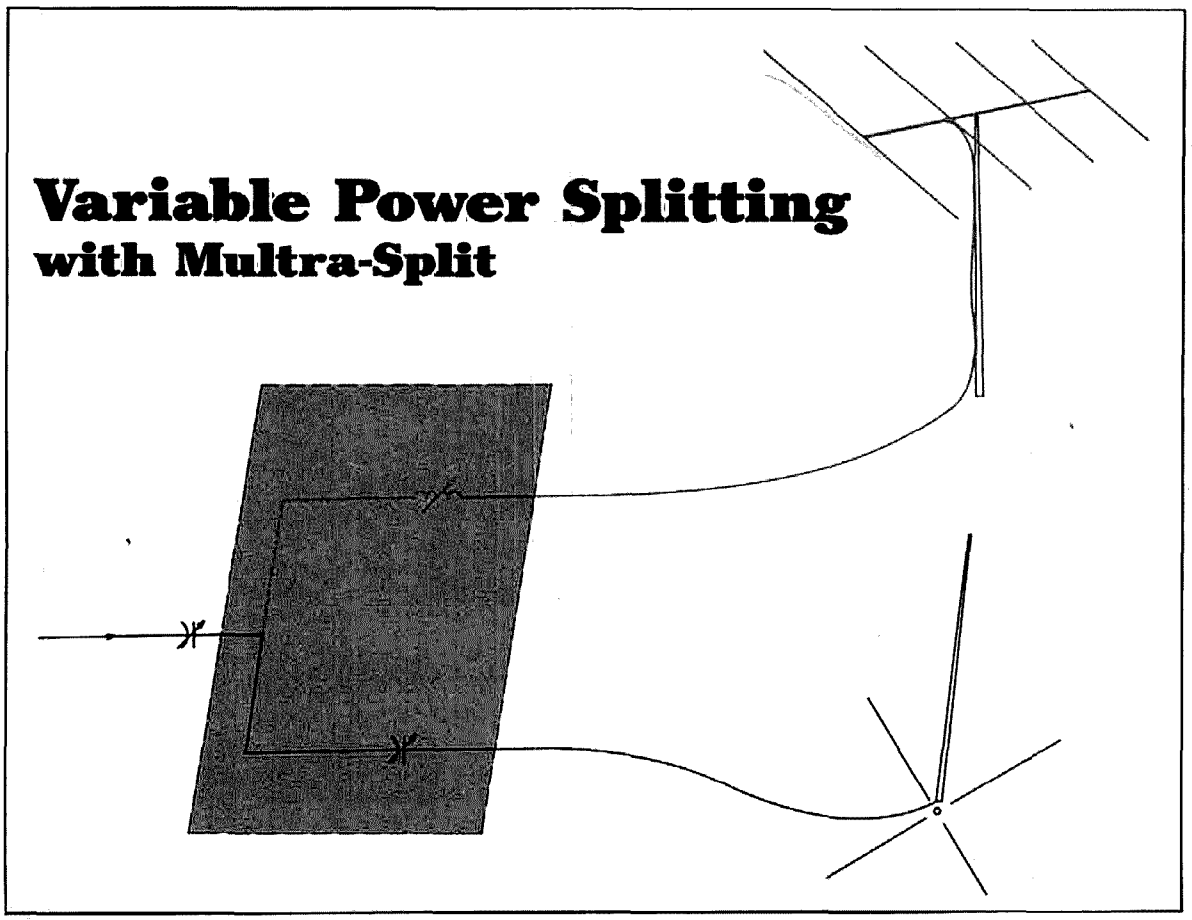
Because of the inherent high-temperature capability of these devices, the thermal consequences of increasing active device density may not be as serious as with semiconductors. It isn't yet clear that the small geometries associated with the "microtubes" translate to correspondingly improved high-frequency performance.

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REFLECTIONS

a wrong-headed solution for someone else's problem

Anyone who carries on a telephone conversation should have a reasonable expectation that the conversation will be private. That's a concept with which it's difficult to quarrel. After all, like motherhood, apple pie, and the Stars and Stripes, privacy is something that the American citizen takes for granted. For those who'd challenge the right to electronic privacy there are severe penalties against wiretapping and (in the Communications Act of 1934) against the misuse or divulgence of any private communication heard over the airwaves.

However, in the view of the few who are utilizing a relatively small portion of the radio frequency spectrum for a very lucrative business — cellular telephone — that's not good enough. After all the time, money, and technical genius they've invested in developing cellular technology, they've suddenly discovered it's possible for others to eavesdrop on what their customers are saying!

So have they harnessed some of their technical genius to solve this terrible problem? Of course not. They've gone to the place where any and all problems are solved easily and at someone else's expense: Washington, D.C.

And what's the solution that Washington has devised? Very simple. Under the terms of bills presently moving through the U.S. House and Senate, the citizens of the United States will no longer have any fundamental right of access to the radio frequency spectrum. As those bills are written, it will become a crime to listen to any radio frequency unless the government specifically decides the frequency is one to which it will permit the general public to listen.

Bizarre? Of course. Absurd? Certainly, but that's precisely the premise of the "Electronic Communications Privacy Act of 1985." To be sure, the Act will include some exemptions, with the first being, as a matter of course, for "broadcasting." From then on, however, the exemptions will get a bit stickier; the Act's avowed purpose is to "prevent" the casual listener, no matter how innocent his intent, from "intercepting" any radio transmission that is not intended for his ears.

The only exemptions to this blanket closing off of the spectrum to casual listening will be those services specifically excluded at their own request or due to the beneficence of the bill's authors. Since WWV's time signals will probably qualify as "broadcasting," you should be able to continue setting your watch with them, but if your receiver draws a bit off 10.000 MHz and you end up using a nearby unidentified CW signal for code practice, you're going to need a good lawyer if you get caught.

"But Amateur Radio will be one of the exempt services," you may object, "so what's our problem?" That's just it. It isn't *our* problem — it's everybody's problem. Have you ever tuned below the broadcast band to ship-to-shore CW around 500 kHz, to aircraft radio ranges, to the "Lowfers" with their 1-watt rigs on 1750 meters, or to the various odd signals below 100 kHz? Forget it. It's going to be illegal. Have you ever tuned between the international broadcast bands to see what else goes on in the HF spectrum? Forget it. Illegal. Like to hear what's happening at the local airport tower, on marine VHF FM, or the various caboose-to-engine railroad frequencies? Forget it. Illegal!

Of course, some of these frequencies might not be illegal; they might just be included in that hallowed list of exemptions. But do you really want some Congressional staffer in Washington to sit down with a large yellow pad and decide for you each and every frequency he'll let you tune to and which will be forbidden, from DC to light? That's the way the bills are written, and that's the way they're expected to work.

Wrong-headed? This proposal certainly is, because it establishes a position that's precisely 180 degrees contrary to the position firmly established for the American people in the earliest days of radio. That position has, from the very beginning, been that the radio frequency spectrum is a public resource, and with only the very minimum regulation necessary, the right of public access to that spectrum should not be denied.

This right really dates from just after World War I, when the U.S. Navy tried to maintain complete control of the airwaves it had been given when the United States entered the war. When the war began, Amateurs were not only off the air, but also required to take down antennas and dismantle receivers. Within weeks after the war's end, a bill was introduced in Congress to continue Navy control of all radio communications in peacetime, but due in part to Amateur opposition, the bill failed. Nonetheless, it wasn't until April, 1919, that the Navy lifted its ban on Amateur receiving. However, the Navy still wouldn't give up its attempt to retain control of all radio until that September, when a Congressional resolution finally forced it to turn radio regulation back to the Department of Commerce, and Amateurs once again regained the right to transmit.

Though in subsequent years the amount of Amateur spectrum has varied from time to time, the right of Amateur Radio to exist — and for the public to listen to whatever it pleased — was never again seriously challenged . . . at least until now. This isn't because the consequences of unlimited public access to the RF spectrum were not understood; these consequences were addressed in the Communications Act of 1934, which made it unlawful to misuse or divulge information derived from listening to private radio communications. In 1934, at least, the framers of the law were wise enough to realize they could not stop casual radio listening without placing a "Kilocycle Cop" in every household.

So what's really wrong-headed about these bills is their basic philosophy. First of all, they take the position that "We, the elite, have the absolute right to decide what you, the populace, can and cannot do with a radio receiver." With this "right," they take from all of us — SWLs, Lowfers, scanner buffs, Amateurs, and even the guy with a short wave or "public safety" band on his kitchen radio — the right to tune a receiver where it pleases us . . . a right that's existed since before anyone even knew how to tune a receiver.

Furthermore, this proposed law is completely unenforceable. It will succeed only in making criminals out of millions of otherwise honest citizens, while the tiny minority that eavesdrops for illicit purposes will continue to do so with impunity, just as it does today. All this repressive and futile effort, simply to "solve" a problem for a narrow-minded special-interest group with a need to cover up its own self-induced shortcoming — lack of communications security — at the expense of the American people.

This makes me mad — mad enough to tell my Senators exactly how I feel about S-1667 and my Representatives what I think of HR-3378. I hope it makes you that mad, too.

Joseph Schroeder, W9JUV

presstop de W9JUV

THERE IS A THREAT TO AMATEUR RADIO IN THE "COMMUNICATIONS PRIVACY ACT OF 1985" in spite of an assumption by some observers that "exemption" from its provisions is sufficient protection. It could hurt recruitment, since many future Amateurs become interested in Amateur Radio through casual listening of the type the new bill would prohibit. Furthermore, the total government control of the radio spectrum that it imposes would almost certainly be applied to Amateurs in the future. (For additional in-depth discussion of the proposal's far-reaching negative effects, see this month's editorial.)

TWO CONTRADICTORY BAND PLANS FOR THE NEW 33 CM (902-928 MHz) BAND are now in use, following adoption of its own plan by the Southern California Repeater and Remote Base Association (SCRRBA). The SCRRBA plan, adopted at an 11/16 meeting, totally ignores the ARRL's VUAC-developed plan -- despite the League's plan having been developed over a period of several years and having been well advertised, both before and after its adoption early last year. For example, the SCRRBA plan not only incorporates different sub-bands, but different FM channel spacing and repeater offsets! Neither plan accommodates the Japanese 903-905 MHz personal radios, which could provide an initial source of equipment to encourage startup activity on the band.

Further Delays In Occupying The Band Will Result from this conflict, with makers hesitating to develop 33 cm equipment compatible with two conflicting standards. As a result, competition from other services or for other uses for this choice 26-MHz wide band will continue to surface. For example, one petitioner has proposed the FCC permit the use of low-power video links -- for example, VCRs to TV sets -- throughout the band. ARRL has filed strong comments against the proposal, designated RM-5193 by the Commission, on the grounds that such joint use would result in unacceptable mutual interference.

TWO NEW RUSSIAN SATELLITES ARE SCHEDULED TO BE LAUNCHED in the very near future, possibly in early February. The two birds, designated RS-9 and RS-10, are expected to be placed in low earth orbit with approximately 120-minute periods. RS-9 will have a single Mode A (2 meters up, 10 meters down) transponder, but RS-10 promises to be the most technologically sophisticated Russian Amateur space effort yet.

Mode K (15 Meters Up, 10 Meters Down) and Mode T (a never-before-used 15 meters up, 2 meters down) transponders are believed to be included in RS-10, along with the usual Mode A transponder. A 2-meter beacon on 145.557 MHz and another on 70 cm (still pending license agreements) will also be included.

RS-10 Is Reported To Have Successfully Completed Checkout and be ready for launch, but RS-9 has had some problems. If problems are resolved in time, the two birds should be launched together; otherwise, RS-10 could go up alone.

Another ISKRA Satellite Launch, In Connection With Refueling of the Russian Salyut 7 Spacecraft, is also thought to be imminent. If so, a Mode K transponder will be included. Since this is an even lower altitude event, the satellite lifetime could be as short as a month or so and the orbital period no more than about 90 minutes.

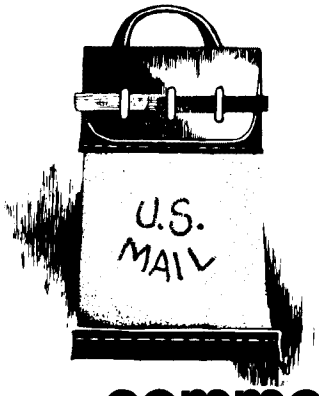
A PETITION FOR A 52-54 MHz "PUBLIC DIGITAL RADIO SERVICE" submitted by Don Stoner, W6TNS, has been designated RM-5241 by the FCC. Stoner's proposal would create a highly automated low-power (1-watt maximum) packet computer-to-computer "hobbyist" radio system, using the top half of 6 meters because of its low usage by Amateurs and unsuitability for other purposes due to the TVI potential. He also suggests there'd be no repetition of the CB chaos of 27 MHz in his service, since he proposes no voice communications -- only data. The FCC released its public notice on RM-5241 on 12/6/85, so the closing date for receipt of Comments was 1/6/86.

THE U.S. AMATEUR POPULATION CREPT TO AN ALL-TIME HIGH of 413,642 individual operators as 1985 drew to a close. The breakdown by license class shows Generals most numerous, with 117,082, followed by Advanced (97,781), Tech (83,387), Novice (77,087), and Extra (38,305). California still has more than twice as many Amateurs as any other state (57,151), with New York (25,897), Florida (24,675), and Texas (24,444) in a close race for second.

Biggest Problem Continues To Be The Novice Class, which is the only license class that's failing to grow. At the end of FCC's fiscal 1983 there were 86,781 Novice licenses in force; a year later that had dropped to 80,461. At presstime the Novice ranks had dwindled by almost another 3400 to 77,087. The continued growth of the higher classes demonstrates the VEC program is working well; the failure is in recruiting new blood.

SPREAD SPECTRUM WILL BE AVAILABLE FOR GENERAL AMATEUR USE ABOVE 420 MHz as of June 1, 1986. Though the new mode may be used for essentially any type of Amateur communication, the rules governing its use are extensive and quite detailed. Potential users should refer to Part 97.71 of the Amateur rules, a new section pertaining only to spread spectrum.

73 MAGAZINE HAS BEEN RETURNED TO ITS FOUNDER, WAYNE GREEN, W2NSD, in a surprise move by CW Communications. CW had acquired 73 as part of the package when they bought Green's computer magazines several years ago, and it was well known in the industry that they've been trying to dispose of it for some time. What is surprising is that Green has returned to the helm, since he'd stated on numerous occasions that he believed Amateur Radio is a dying hobby and he no longer had any serious interest in it.



comments

odd antennas

Dear **HR:**

Those of us who have been hams for 30 years but are not radio engineers sometimes wonder about certain unused antenna designs. Perhaps the expertise of the readership of *ham radio* can enlighten us on these peculiar issues.

For instance, every antenna handbook I own describes the advantages in gain of 5/8-wave verticals *and* 1.28 wave dipoles, the 1.28 wave dipole having a 3-dB gain over a half-wave dipole (see the Extended Double Zepp, fig. **1**). Why aren't more antennas designed using the 1.28-wave element, especially at frequencies above 144 MHz?

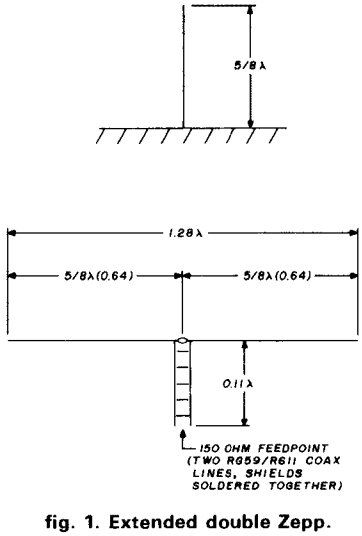


fig. 1. Extended double Zepp.

Every good handbook describes the Windom 14 percent off-center-fed antenna (fig. **2A**), normally usable at 1/2, 1, and 2-wave frequencies — three bands. Why aren't multi-band dipoles (fig. 2**B**) tubular beams (fig. **2C**) and quads (fig. **2D**) designed for 3-bands per element 14 percent off-center-feed?

And since the Windom 600-ohm feedpoint occurs at 14 percent off-center (or *64* percent from one end) and the high-gain 5/8-wave vertical is a 62.5 percent feed point, why won't a bent-Windom vertical, fed 64 percent up, with a 36 percent horizontal section, produce a near 5/8-wave superior

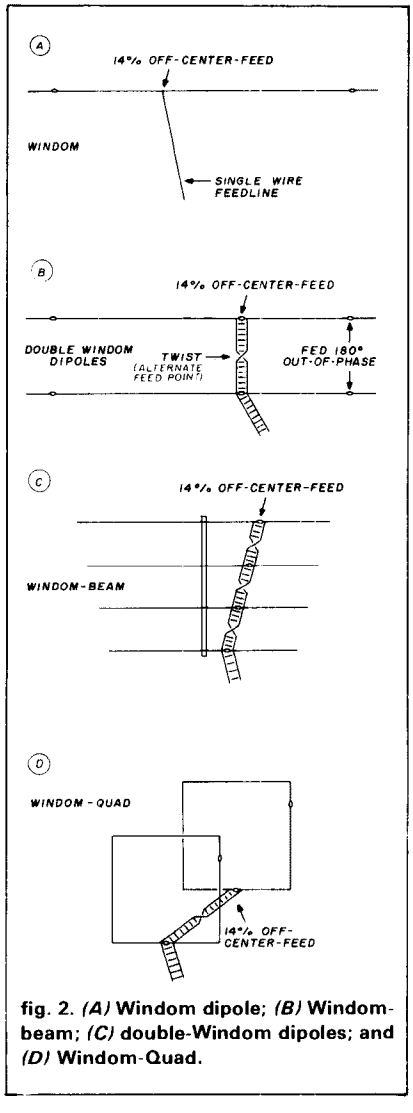


fig. 2. *(A)* **Windom dipole;** *(B)* **Windom-beam;** *(C)* **double-Windom dipoles; and** *(D)* **Windom-Quad.**

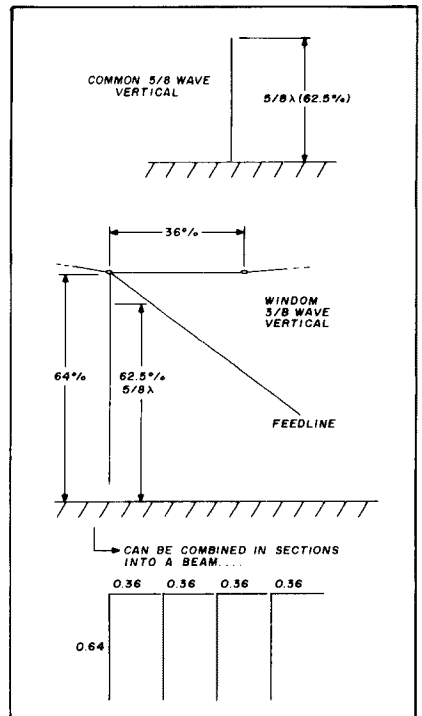


fig. 3. Windom dipole.

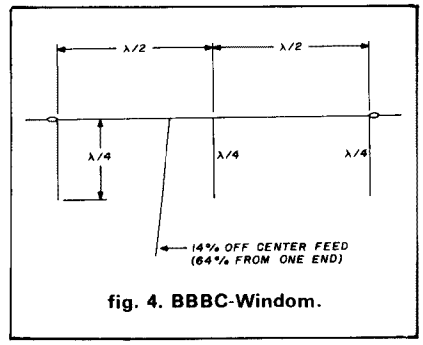


fig. 4. BBBC-Windom.

signal plus a signal on two other bands (fig. 3)?

Since the BBBC (Broadside Bidirectional Bobtail Curtain) (fig. 4) antenna is excellent on one band for DX, why can't it be fed at the 600-ohm off-center Windom feedpoint to make it a multiband antenna?

No doubt there are very good reasons why these antennas do not exist, but a good, logical answer would be gratifying to those of us empirical tinkerers who are theory-deficient.

C.N. Francis, WØMBP
Silsbee, Texas

achieve polarization diversity through variable power splitting

Multra-split circuit increases signal strength and extends transmission times

The reception of high frequency (3-30 MHz) signals propagated via the ionosphere is characterized at times by irregular and deep fading (indeed, in such places as the auroral zone, by long periods of excessive absorption resulting in communications blackout). This is related to varying amounts of energy in the vertical and horizontal components of the radio wave. A device at the transmitter that takes account of this condition will enhance communications. Power is divided between two transmitting antennas in such a manner that ratios of from 1:1 to 10:1 can be effected without mismatch. In this manner, vertical and horizontal components of the transmission pattern are compensated to prevent fading at the receiving location utilizing radiation angle diversity. A characteristic of the entire propagation phenomenon during altitude changes of the ionosphere is that the horizontal and vertical components never fade simultaneously. Power splitting takes advantage of this characteristic when other parameters of the system have already been optimized. In this way, communications on a particular band may be extended for up to several hours using this technique.

It is generally accepted that an HF antenna with a low angle of radiation is capable of sustaining usable communications for additional periods as long as two to three hours, during the ionosphere transitional periods.

It is also recognized that during this transitional time some disruption of communications circuits takes place. HF stations normally change their frequencies to re-establish communications. Therefore, any reasonable means that will allow a communicator to extend the transmission period and/or improve the reliability of communications during a transition period, without changing bands, is worth investigating.

During the last 15 years, I have conducted a series of limited experiments on 15 and 20 meters using a transceiver and a pair of antennas. One was a six-element horizontal trap-beam, and the other a vertical antenna. Power from the transmitter was divided between the two antennas by means of a network called *Multra-split*. The Multra-split network allowed the power to be divided anywhere from 1:1 to 10:1 without mismatch at the transmitter.

In practice, the antenna with the highest gain is fed the highest power, and generally the higher gain antenna is horizontally polarized. However, depending upon the band of transmission and time of day, it can be more favorable to feed higher power to the vertical antenna.

sky-wave propagation[1]

At high frequencies, long-distance communication is possible only via sky-wave propagation, whereby radio waves are refracted back to earth from one of the surrounding ionized layers. These layers are the

By John H. Mullaney, W3NGJ, 9049 Shady Grove Court, Gaithersburg, Maryland 20877.

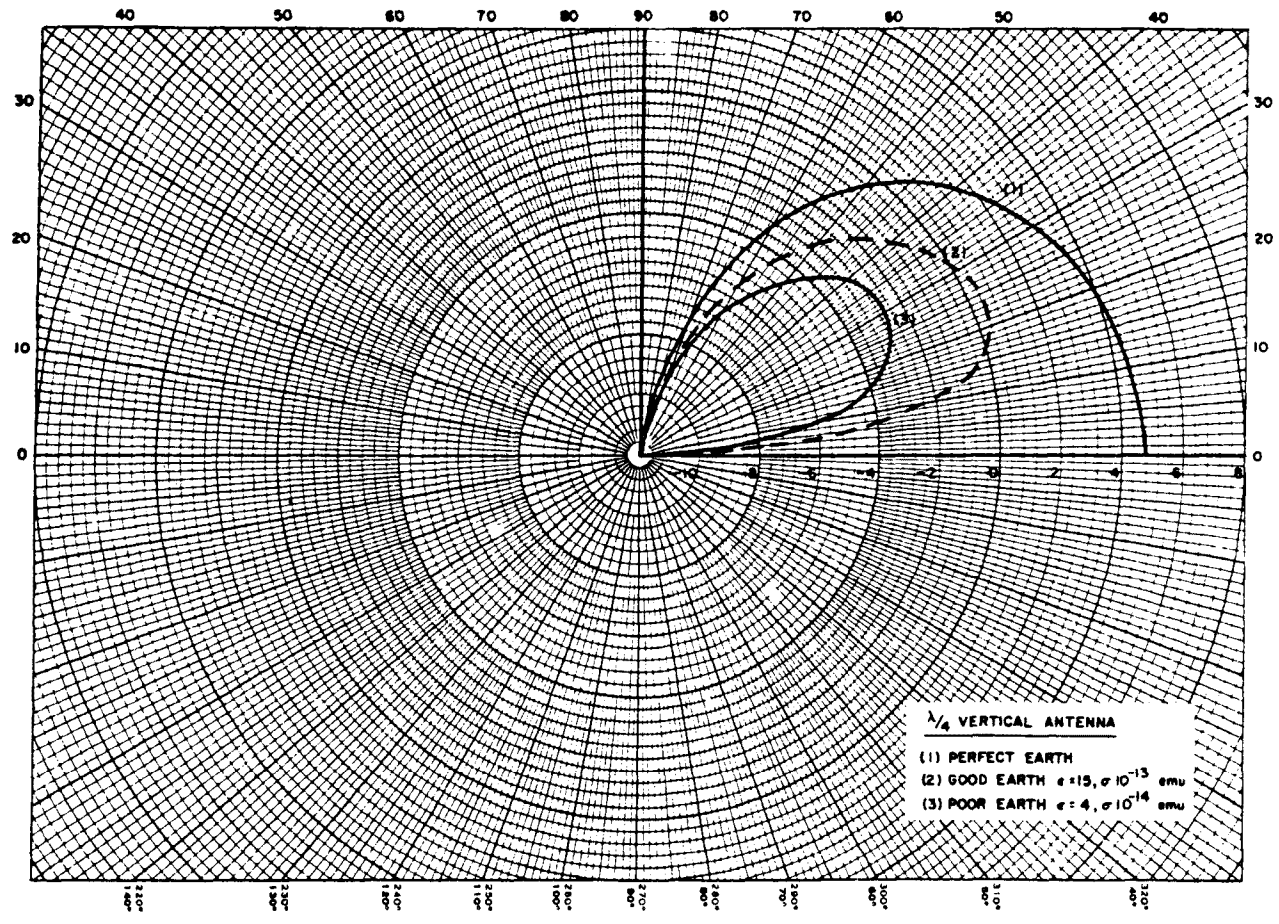


fig. 1. Vertical radiation pattern over earth.

result of ionization of neutral gas in the atmosphere by ultraviolet radiation from the sun.

Because of the variation in the structure of the atmosphere with height and the selective absorption by the various gases of the radiation from the sun, the ionization tends to have several regions of maximum electron density at various heights. These regions are the E-layer, at an average height of 70 miles (113 km), the F1 layer, at about 140 miles (225 km), and the F2 layer, at approximately 250 miles (402 km). Below the E-layer, during the day, is another ionized region at an average height of 35 miles (56 km). This is called the D-region. The D-region accounts for daytime absorption of medium-frequency radio waves. This is the principal reason 80 and 160 are considered night-time bands. These layers, as might be expected (since radiation from the sun is the principal ionizing agent), are present during daylight. At night the D-region is not present, the F1 layer merges with the F2 layer at an average altitude of approximately 175 miles (282 km), and the E-layer essentially disappears.

polarization

Antennas may be classified as linear elements, apertures, arrays, or traveling wave types. For our discussion, and considering that we are interested in Amateur communications in the 2-30 MHz range, we will deal only with linear and array types.

Two field components are present on an antenna: the magnetic and electric fields. Radio waves, like light waves, can have a definite polarization. Radio wave polarization is determined by the orientation of the electric field component, which in turn is determined by the orientation of the elements themselves.

On a linear antenna or an array of elements, the magnetic field component is always at right angles to the radiator, whereas the electric field component is always in the same plane as the radiator. This means that a vertical antenna or radiator will transmit a vertically polarized wave and a horizontal radiator will transmit a horizontally polarized wave.

Most of the common optical effects experienced

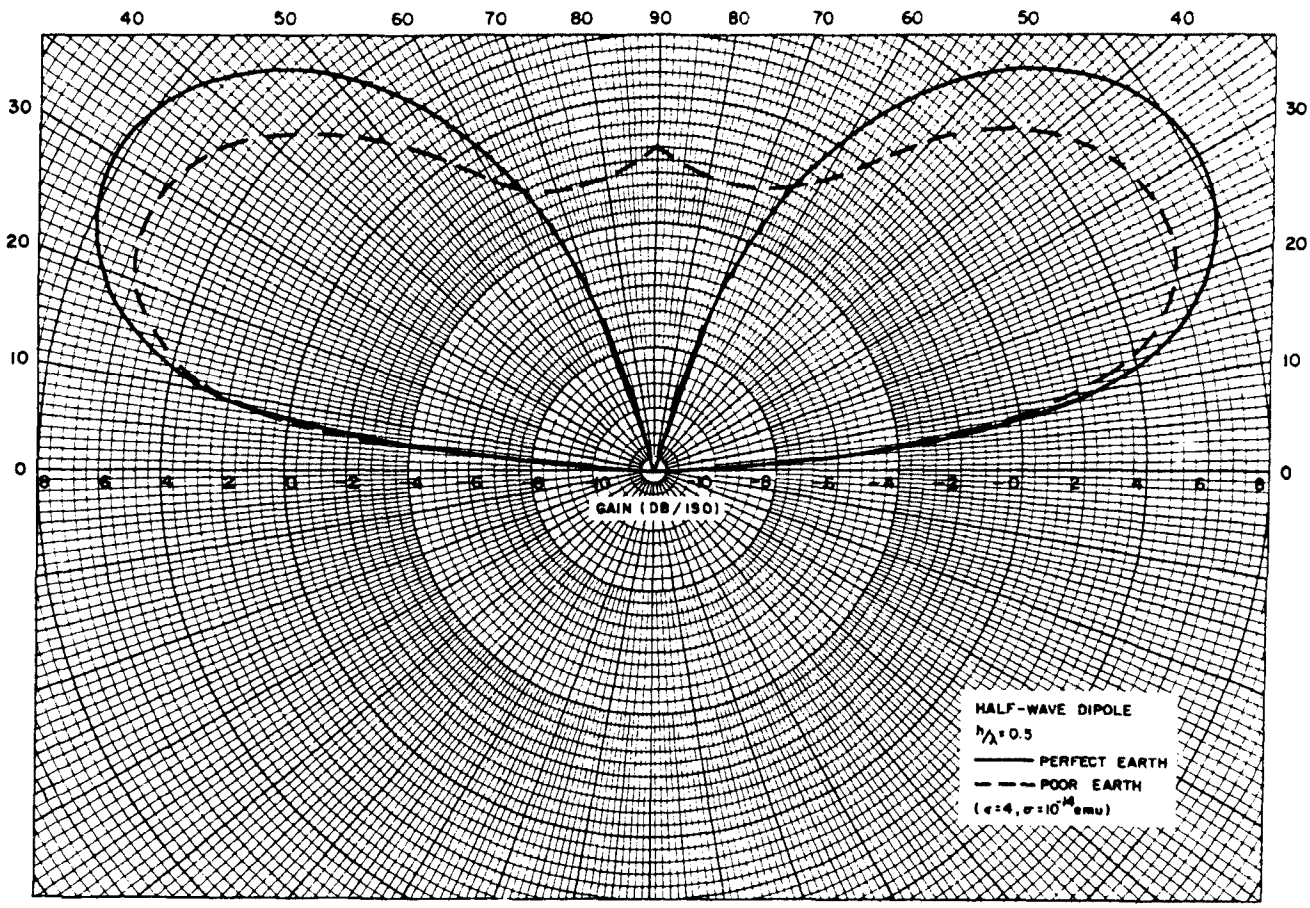


fig. 2. Vertical radiation pattern for horizontal half wave dipole at 180 degrees above earth.

with polarized light also occur between pairs of antennas having different polarizations, when the medium (such as free space) between the antennas does not affect polarization. A linear antenna used for receiving responds most strongly to another antenna with the same polarization, and not at all to one polarized at 90 degrees to the first antenna.

Vertically polarized waves radiated by a vertical antenna over imperfect ground are elliptically polarized because of the presence of a small electric field along the ground in space quadrature and the different time phase from the major electrical component.

For high-frequency ionospheric layer propagation, the received signal is usually elliptically polarized, even when the transmitting antenna produces linear polarization. This condition arises from the effect of the earth's magnetic field on the ionized layers; it splits the incident linearly polarized signal into an ordinary and extraordinary wave. The two waves travel at different velocities and experience different polarization rotations. The resultant received signal is elliptically polarized; consequently, depending on the orientation of the receiving antenna and its height above ground, the received signal may not be usable. The use, therefore, of a vertically and horizontally polarized antenna with a special combining network (such as Multra-split) may provide improved communications not only from a signal strength standpoint, but also over an extended period of time.

The effectiveness of this method of propagation is directly affected by the radiation axis or radiation angle of the antennas used. Whether an antenna is vertically or horizontally excited, its vertical radiation pattern shape is a function of its electrical and physical heights above the ground.

Figures 1 and 2 show the vertical radiation pattern for a horizontal and vertical antenna.

Note that the vertical shape of the elevation pattern is a function of the earth's conductivity. The lower the conductivity, the more pronounced the distortion to the vertical pattern (high angle signals are hardly attenuated). Either antenna can be selected for the high power side, depending upon transmission path conditions.

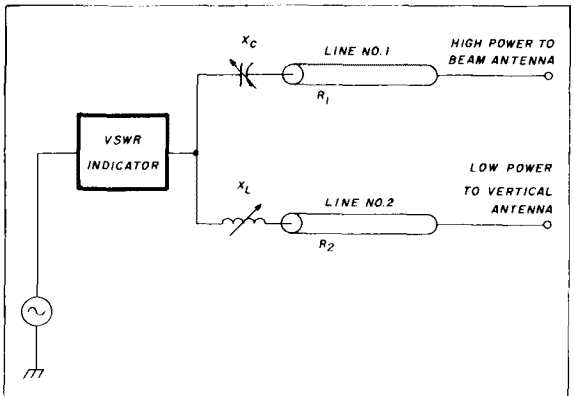


fig. 3. Low power antenna connected to inductive branch.

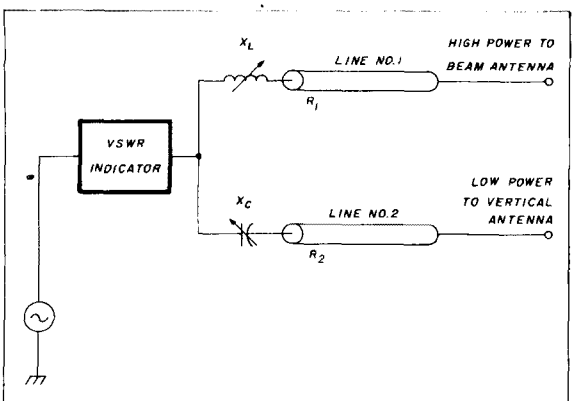


fig. 4. High power antenna connected to inductive branch.

One way to enhance reception and transmission in a varying polarization environment is to use a combiner/power divider network that responds more favorably to the dominant polarization.[2,3] A transmitting type circuit that provides this function (Multra-split) is shown in **fig. 3.** In this case, two antennas — one horizontally polarized, the other vertically polarized — are fed by a network that splits the power as required. This configuration provides more power to the beam antenna. If the polarization that propagates best is vertical, then the **fig. 4** configuration is more appropriate.

circuit operation

The Multra-split circuit is basically a simple power divider. There are two equal loads (50 or 70 ohms), R1, and R2, which represent the transmission line resistance for antennas 1 and 2. Reactances X_C and X_L are of opposite sign and their values are uniquely determined by the power-division ratio and input impedance desired. In this case, the input impedance is selected to match the output of the transmitter (50 or 70 ohms). For our discussion, we will assume the use of 50-ohm transmission lines. The power division is determined by the transmission path conditions. Generally, with a high-gain horizontal beam and a vertical whip, the higher power is fed to the beam and just enough power is fed to the vertical antenna to obtain the diversity effect.

The power division can be expressed in terms of a factor M′, as follows:

$$M' = \frac{P\ (hi)}{P\ (low)} \qquad (1)$$

where

P (hi) = power to fed to Line 1,
P (low) = power fed to Line 2.

The value of reactance for the high-power leg (Line 1)

is:

$$X = \pm \frac{R1}{M'} \qquad (2)$$

where

X = value of positive or negative reactance in ohms,
R1 = Line 1 resistance in ohms,
M′ = power-division factor defined above.

The value of the reactance for the low-power side (Line 2)

is:

$$X = \pm M' \cdot R2 \qquad (3)$$

where

X = value of positive or negative reactance in ohms,
M′ = power-division factor,
R2 = line 2 resistance in ohms (must be equal to Line 1 resistance).

It will be demonstrated later that the input impedance will always be equal to the transmission-line impedances, provided certain design requirements are met. This is the same as saying that, under the required conditions, the feedpoint resistance will always remain equal to the load resistance, regardless of the power division between the two branches.

practical application

Assume a station desired to divide its power on a 90/10 basis; that is, 90 percent in line 1 and 10 percent in line 2. The known parameters are:

f (frequency) = 14.250 MHz.
P (transmitter output) = 500 watts.
R (transmitter output impedance) = 50 ohms.
I (transmitter output current for 500 watts)

$$I = \sqrt{\frac{P}{R}} = \sqrt{\frac{500}{50}} = 3.16 \text{ amperes.} \quad (4)$$

Current in line 1 for 450 watts:

$$I = \sqrt{\frac{450}{50}} = 3.0 \text{ amperes} \quad (5)$$

Current in line 2 for 50 watts:

$$I = \sqrt{\frac{50}{50}} = 1.0 \text{ ampreres} \quad (6)$$

Power division factor:

$$M' = \sqrt{\frac{P\ (hi)}{P\ (low)}} = \sqrt{\frac{450}{50}} = 3 \quad (7)$$

An inductor is selected arbitrarily for the high power branch. The reactance is:

$$X_L = + \frac{R1}{M'} = \frac{50}{3} = 16.7 \text{ ohms.} \quad (8)$$

Because we selected an inductor for the high power side, a capacitor must be used in the low power branch. Its reactance is:

$$X_C = M' \cdot R1 = 3 \times 50 = 150 \text{ ohms.} \quad (9)$$

and since it's a capacitive reactance = −150 ohms. The magnitude of X_L and X_C have now been determined. The inductance and capacitance can be readily found. The inductance is:

$$L = \frac{0.159 X_L}{f} \quad (10)$$

where

L = inductance in microhenries
X_L = inductive reactance in ohms,
f = frequency in MHz.

solving for L

$$L = \frac{0.159 \cdot 16.7}{14.25} = 0.186\ \mu H \quad (11)$$

The capacitance is determined by:

$$C = \frac{159 \times 10^3}{f \times X_C} \quad (12)$$

where

C = capacitance in picofarads,
X_C = capacitive reactance in ohms,
f = frequency in MHz.

Then:

$$C = \frac{159 \times 10^3}{14.25 \times 150} = 74.39\ pF. \quad (13)$$

The circuit can now be redrawn as in **fig. 5.**

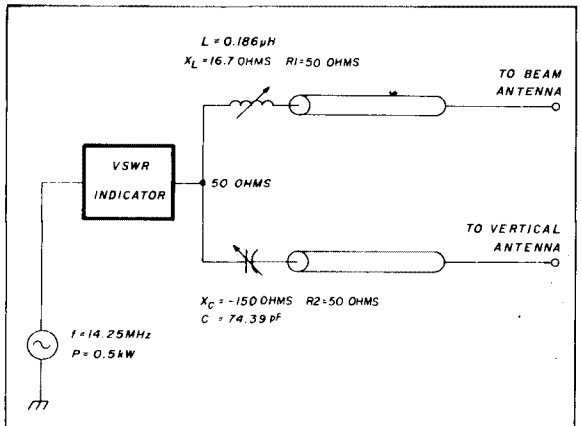


fig. 5. Multra-Split network with matching resistance loads.

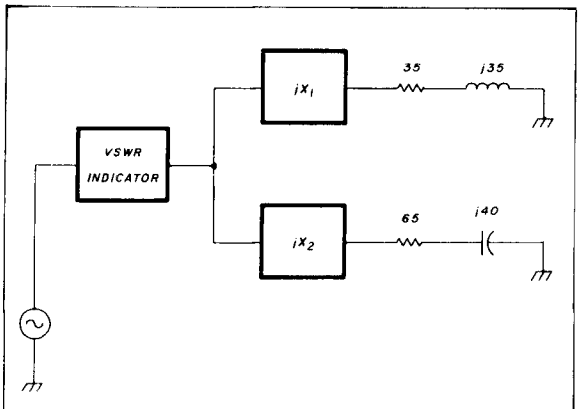


fig. 6. Equivalent circuit for determining components to offset mismatch.

circuit variations

Combinations of X_L and X_C can be selected to accomplish different power splits.

In many cases, it's more practical to use variable capacitors and inductors to obtain the exact power ratio; in others, a fixed capacitor can be made equivalent to a variable unit by using a variable or tapped coil in series with a capacitor.

If you want to operate a transmitter over several bands of frequencies, provisions should be made to switch in different size components. The size and rating of components from a voltage and current standpoint are determined in the usual manner for any network.

It should be noted that the circuit works on the principle that both lines are equal in magnitude and nonreactive. In the event that one or both of the lines are mismatched, it's necessary to adjust the X_C or X_L or

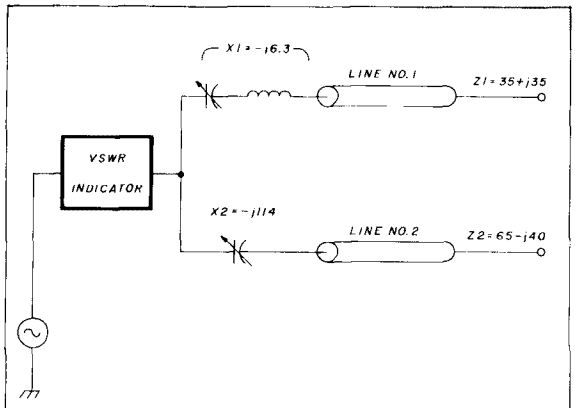


fig. 7. Series network.

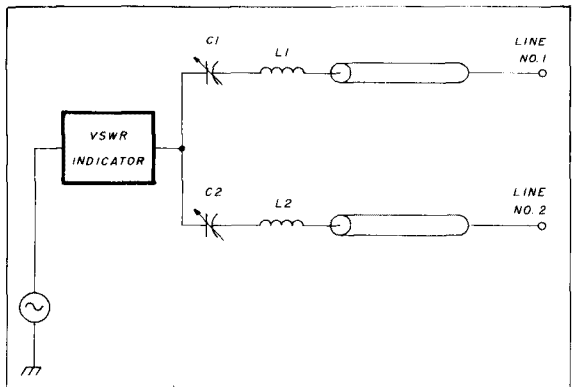


fig. 8. Typical network permitting tuning both arms for plus or minus reactance.

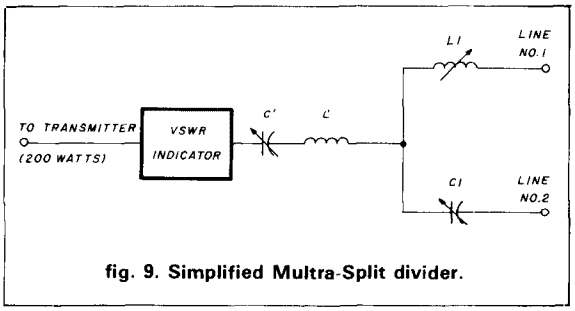


fig. 9. Simplified Multra-Split divider.

both to obtain a 50-ohm input for the transmitter. A VSWR indicator at the transmitter output or network input will indicate a match condition.

Where either or both antennas have a large mismatch, the method for determining the combining components becomes somewhat more complex because we have to deal with four variables; that is, two resistances and two reactances. The problem, however, is not too difficult to solve.

Example Assume:

Transmission line 1 = 35 +j35.

Transmission line 2 = 65 −j40

Power distribution M′ = 3.00 (90 percent/10 percent).

Tramsmitter impedance 50 ohms.

The equivalent circuit can be represented as in **fig. 6**.

The values for jX1 and jX2 are of opposite sign and are determined as follows:

$$X1 = \left[\frac{b^2d - 2ea + (b^2d - 2ea)^2 + 4e^2(b^2c - a^2)^{1/2}}{2(b^2c - a^2)}\right]^{1/2} \quad \textbf{(14)}$$

where

$X1$ = *net reactance for line 1 (the load reactance),*

$$a = R0\text{-}R2 + M'^2\left(\frac{R2}{R1}R_0 - R_1\right)$$

$$b = \frac{2M'R0}{R1 + R2}$$

$$c = R2/R1$$

$$d = R2\,R1\left[\frac{R2^2}{M'^2} - R1^2\right]$$

$$e = R1R2 - R0R1 - R0R2 + M'^2\left(\frac{R2}{R1}\right)\left(\frac{R2^2}{M'^2} - R1^2\right)\left(\frac{R0 - R1}{R1 + R2}\right),$$

R_O = *transmitter resistance = 50 ohms*

Substituting the known parameters, X1 = j 28.7 ohms.

Remember that X1 represents the net reactancce required for branch one; hence, because we already have j 35 in the circuit we must wash out j 6.3 ohms (35-28.7) = 6.3) to obtain the required net reactance. This is best accomplished by a series network adjusted to have a reactance of −j 6.3 ohms in series with line 1.

X2 is determined from (and will be a negative reactance because X1 is positive).

$$X2 = M'\left[\frac{R2}{R1}R1^2 + X1^2 - \frac{R2^2}{M'^2}\right]^{1/2} \quad \textbf{(15)}$$

Substitute in **Eqn 15** X2 = −j 154 ohms. Inasmuch as the reactance for line 2 is −j 40, we have to add a minus of −j 154 as required by **Eqn 15**. We obtain this value by using a capacitor which is adjusted to −j 114 ohms.

We can now redraw our circuit as in **fig. 7**. In practice we should use a circuit that has series networks in both arms because we can adjust these for either

plus or minus reactance, depending upon which side of resonance we are tuned. **Fig. 8** is a typical circuit.

Due to the fact that we are working with mismatched transmission lines, the maximum current and voltages at the input of the lines are determined as follows:

and

$$E_{MAX} = [P \cdot Z_o \times VSWR]^{1/2} \quad \textbf{(16)}$$

$$I_{MAX} = \left[\frac{P \times VSWR}{Z_o}\right]^{1/2} \quad \textbf{(17)}$$

where

P = power in watts

VSWR = voltage standing wave ratio,

Z_0 = characteristic impedance of line.

To demonstrate the effect of a mismatch or VSWR vs. a pair of matched lines (**Eqns 1, 14,** and **15**) were used to determine the net reactances required for X1 and X2 for the condition where R1 = R2 = 50 ohms,

table 1.

P1 percent	Matched condition Whore R1 = R2 = 50 ohms X1 ohms	X2 ohms
10	150.00	16.67
20.00	100.00	25.00
30.00	76.38	32.73
40.00	61.24	40.82
50.00	50.00	50.00
60.00	40.82	61.24
70.00	32.73	76.38
80.00	25.00	100.00
90.00	16.67	150.00

table 2.

P1 percent	Mismatched condition Where R1 = 35 ohms and R2 = 65 ohms X1, ohms	X2 ohms
10	192.81	12.59
20.00	126.32	18.24
30.00	97.08	28.09
40.00	78.98	38.08
50.00	65.86	48.86
60.00	55.36	61.51
70.00	46.25	77.93
80.00	37.68	102.73
90.00	28.71	153.85

table 3.

P1 percent	C1 (PF)3	L1(μH)
10.00	58	0.141
20.00	88	0.204
30.00	115	0.313
40.00	141	0.425
50.00	169	0.545
60.00	202	0.686
70.00	241	0.869
80.00	296	1.146
90.00	389	1.717

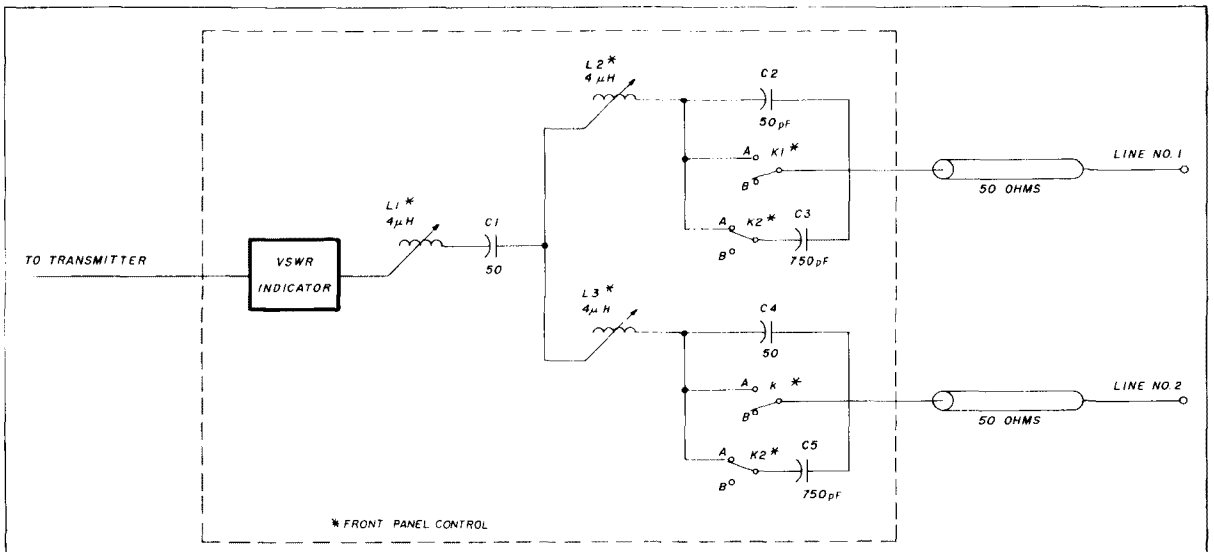


fig. 10. Three band Multra-Split power divider.

and R1 = 35 ohms and R2 = 65 ohms. **Table 1** represents the matched condition, whereas **table 2** is for the mismatched condition.

The differences for identical powers is apparent from a study of the tables. In practical applications, series networks in both legs are used with components large enough to handle any mismatch.

typical multra-split circuits

Figure 9 illustrates a typical stripped-down version of a Multra-Split power divider for one Amateur band.

C' L' is a series circuit that can be adjusted either side of resonance to wash out any residual reactance due to mismatch on the transmission lines, while L1 and C1 are the components used to control the percentage of power to each antenna.

Table 3 furnishes typical component values for C1 and for power division from 10 percent to 90 percent where one antenna is a beam and the other a vertical. This divider is for 20 meters with the center frequency being 14.25 MHz. We are using the inductive branch for high power and assume both lines could be mismatched.

C1 would be a variable capacitor of approximately 25-400 picoFarad, while L1 would be a variable inductor of approximately 2.5 μH. The voltage and current ratings of the components would be determined by the transmitter power and VSWR.

The series circuit L' C' would consist of a fixed inductor of 4 μH, while a variable capacitor of 25-400 picoFarad would be used for C'.

Figure 10 illustrates another Multra-Split power divider, used by the author, to cover the bands of 10, 15, and 20 meters. It will handle 1 kW of power.

conclusions

The Multra-Split network provides a simple method of dividing power between two loads or antennas of identical or dissimilar polarization radiation angles. It can be used to improve HF communications because it works on the principle of taking advantage of radiation angle diversity. Electrically, the principle is straightforward and it should provide a practical economical method for power splitting.

acknowledgement

Thanks to George Jacobs, W3ASK, for his valuable suggestions and help in preparing this article.

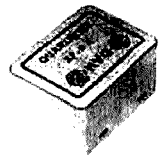
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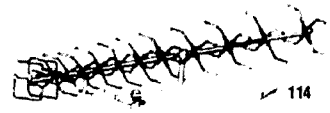
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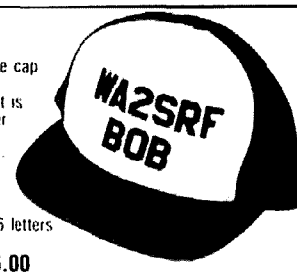
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cylindrical feedhorns revisited

Another look at this popular method

The subject of feedhorn design is obviously of interest to Amateurs and experimenters everywhere; nine years after my first *ham radio* article on this topic, and almost three years after the second, was published, inquiries are still being received.[1,2]

The concept of the open-ended waveguide for use as an antenna or primary feed for a parabolic reflector was covered in the first article; the second article addressed impedance matching to space. Because of limited test pattern facilities, the matter of pattern shape was not thoroughly discussed.

It was only recently that I had an opportunity to make radiation patterns of the cylindrical feedhorn on a professional pattern range with commercial equipment. These tests were made at frequencies between 3.7 and 4.2 GHz because the feedhorns were designed for that frequency range. The results, however, apply equally well on Amateur frequency bands such as 1296, 2304, 5700 and above, wherever parabolic reflectors would normally be used.

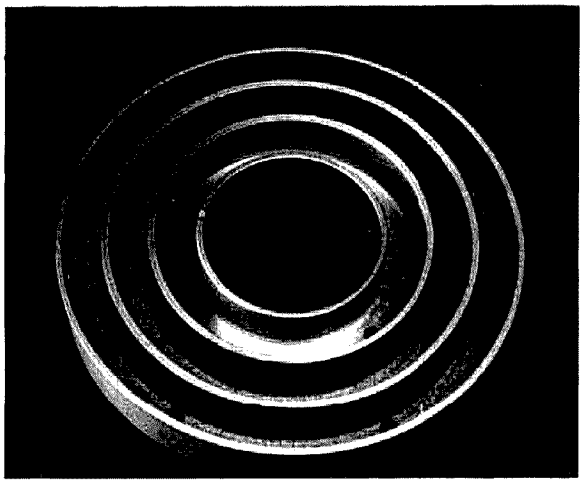


fig. 1. Photograph of typical scalar choke.

A scalar feedhorn is generally defined as one employing an RF "choke" mounted on the outside of the horn near the open end. Commercial feedhorns often employ chokes consisting of four to six concentric rings, either diecast or milled from a block of aluminum. The purpose is to choke off currents that would otherwise flow on the outside surface of the horn, thereby producing unwanted radiation and power loss. A photograph of a typical scalar feedhorn choke is shown in **fig. 1**.

A scalar choke can be fabricated for use at Amateur frequencies by forming copper strips into concentric circles and soldering them onto a brass washer. The height of the strips should be slightly less than a quarter wavelength, spaced approximately 0.2 wavelengths. The latter is not critical, but care should be taken to assure that the circles are concentric.

The choke shown in the second article was a large flat brass washer soldered to a short cylinder that slid over the outside of the horn for positioning. This type of choke is very effective in improving horn efficiency, and is much easier to fabricate than the scalar choke.

Radiation patterns recently made show that the pattern shape using the flat choke is slightly asymmetrical at some frequencies, causing a few degrees of "squint." That is, the axis of the beam tends to be off the horn axis, and this tends to vary with frequency, as shown in **fig. 2**. The shape of the pattern, and in particular its symmetry, is highly important in terms of overall antenna efficiency when a feedhorn

By Norman J. Foot, WA9HUV, 293 East Madison Avenue, Elmhurst, Illinois 60126

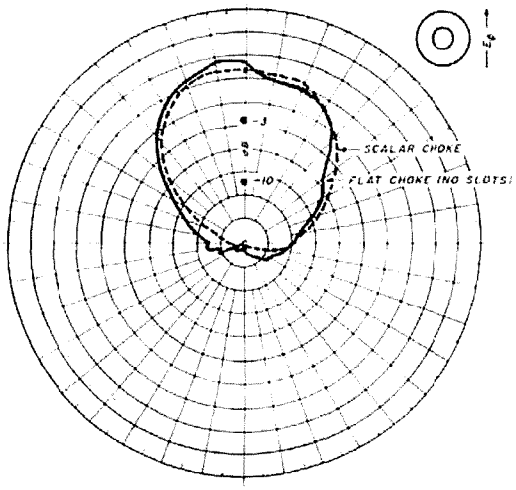


fig. 2. Scalar and flat choke pattern comparison. The flat choke without slots shows a skew off boresight.

is used together with a parabolic reflector, especially when used over an appreciable frequency band.

overcoming squint

The addition of a single set of resonant slots cut into the choke overcomes the squint problem. The slots act as RF chokes in a manner similar to the concentric rings of the scalar choke. With the addition of the slots, squint is removed and patterns are smooth and axially symmetrical. **Figure 3** is a photograph of a flat choke with slots. Thus, while the addition of a flat choke to the cylindrical feedhorn produces a good match to free space, the slots choke off residual currents that would otherwise flow on the outside of the horn and result in squint. **Figure 4** and **5** compare the patterns of a feedhorn employing a scalar choke with the same horn using a slotted choke. Boresight gains are identical, and 3 and 10-dB beamwidths are for all practical purposes identical. **Figure** 4 also shows what can happen when the choke is not used.

slot orientation function of polarization

In operation, the slots are oriented along the direction of polarization. When used with vertical polarization, the slots are located one above the other. Rectangular slots cut in a conducting ground plane are sometimes used as radiating antennas; they are normally excited by connecting a coaxial cable across the small dimension. The resulting RF current tends to flow parallel to the large dimension of the slot. The slot performs like a dipole except that the directions of voltage and current are reversed. The long dimension is approximately equal to a half-wavelength.[3]

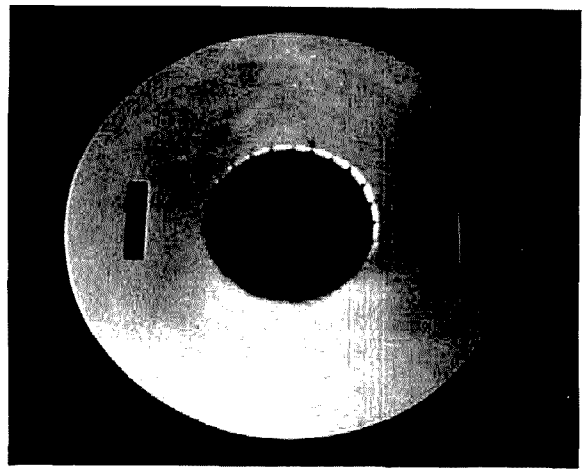


fig. 3. Photograph of slotted choke.

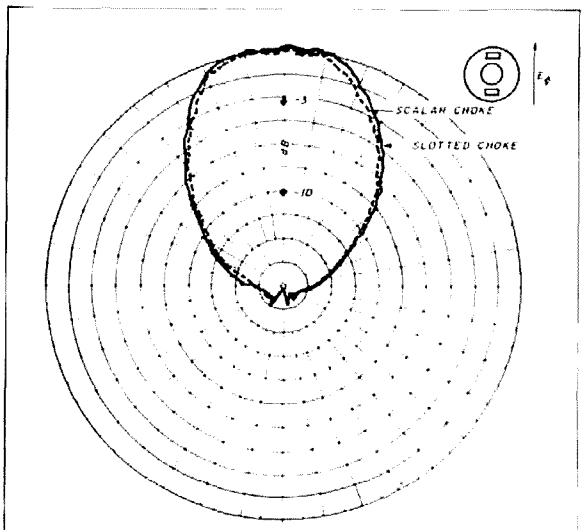


fig. 4. Patterns of vertically-polarized slotted and scalar chokes.

When currents flowing radially outward on the flat choke encounter slots oriented in the wrong direction (i.e., normal to the direction of slot current flow), the current tends to be inhibited and the slots act like RF chokes. Choking action is most effective when the slots are resonant at the RF operating frequency. The slots for use at 4 GHz were cut 1.4 inch long and 0.25 inch (3.17 x 0.63 cm) wide. The slot width is not critical as long as it is much smaller than a wavelength.

If the antenna system uses both horizontal and vertical polarization, such as is the case with TVROs two sets of slots should be used to avoid having to rotate the choke 90 degrees to change polarization. A choke of this kind is illustrated in **fig. 6**. Not shown are a

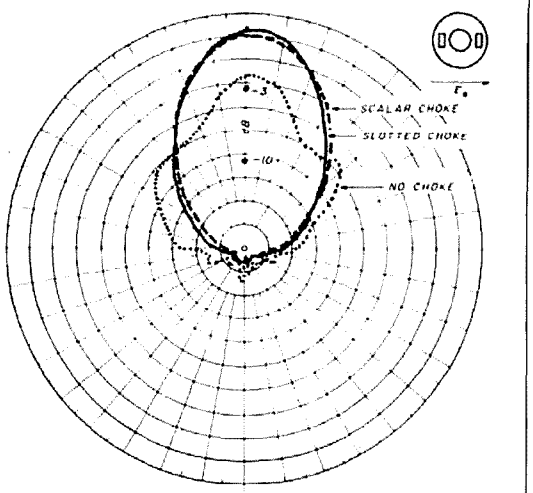


fig. 5. Patterns of horizontally-polarized slotted and scalar chokes.

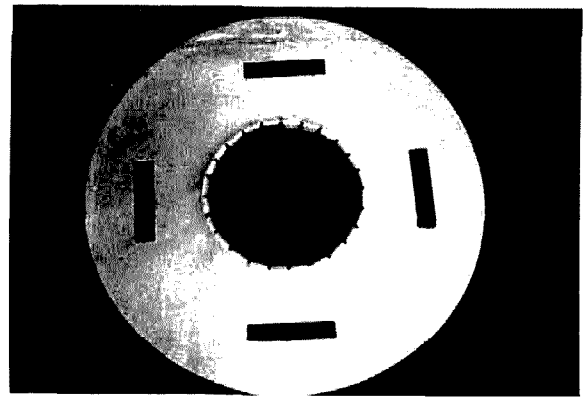


fig. 6. Double slotted choke.

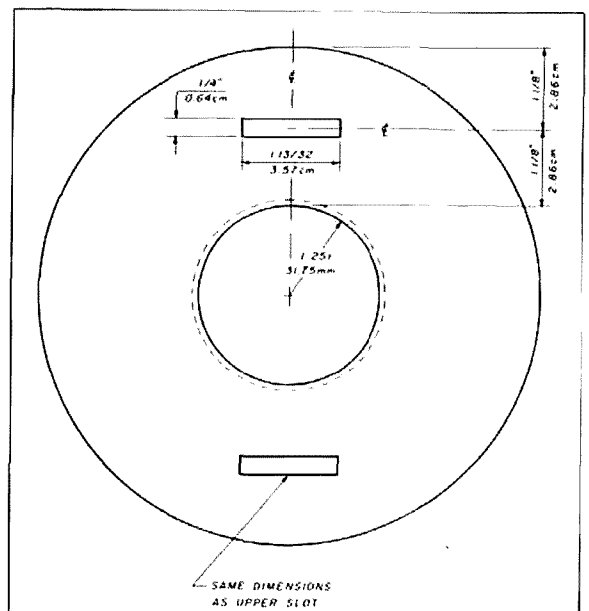


fig. 7. Drawing of slotted choke used in pattern tests.

pair of orthogonally mounted monopoles inside the horn, each of which responds to its respective polarization. Two sets of slots would probably work well with circular or cross-polarization as well, although this has not been confirmed by test.

choke dimensions

The 4 GHz choke shown in **fig. 3** has a 7-inch (17.8 cm) OD and a 2-1/2 inch (6.35 cm) diameter hole at its center which slides over the 2-1/2 inch (6.35 cm) OD horn. The inside diameter of the horn is 2-1/4 inches (5.72 cm) and it is 4 inches (10.16 cm) long. A mechanical drawing of this choke is illustrated in **fig. 7**. In the pattern tests, the optimum location for the slotted choke was found to be in the plane of the open end of the horn. Pushing the choke forward had very little effect on the gain or patterns, but gain falls off somewhat as the choke is moved to a position behind the horn opening.

Choke dimensions can be scaled for operation at other frequencies. The 2304 MHz choke should be 12 inches (30.48 cm) OD and the center hole should be slide-fit with the 3.9 inch (9.91 cm) diameter of the horn. One-pound coffee can make ideal horns for 2304 MHz, since their ID is slightly less than 4 inches (10.16 cm). The choke slots should be 2-1/2 inches (6.35 cm) long and about 3/8 inch (0.95 cm) wide.

At 1296 MHz, the choke diameter becomes relatively large. If scaled from **fig. 7**, it would be 21 inches (53.3 cm). To avoid excessive shadowing, especially when using a relatively small dish, a choke diameter as small as 18 inches (45.72 cm) provides reasonably good performance at 1296 MHz.

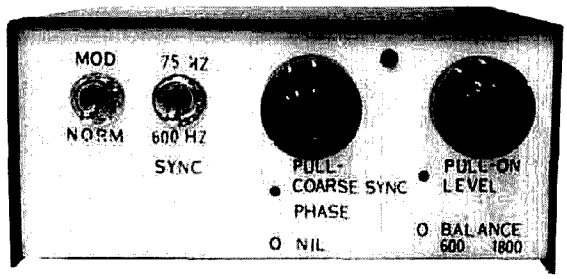
A 7/8-inch (2.22 cm) ID brass tube makes an ideal horn for the 10.0-10.5 GHz band. The choke should be 2.7 inches (6.86 cm) OD with 9/16 x 1/8 inch (1.43 x 0.32 cm) slots.

Reader inquiries may be addressed to Footronics Engineering Inc., 293 East Madison Avenue, Elmhurst, Illinois 60126, in care of WA9HUV. Please enclose an SASE.

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ham radio



two-tone signal generator

Check linearity of SSB transmitters with an easy-to-read oscilloscope display

This article describes a two-tone test generator for checking SSB transmitter linearity. In addition to supplying two low-distortion audio tones to the microphone input of a transmitter, it has three additional features. The first permits the generation of a "bow tie" display on an oscilloscope in which linearity shows up as a straight line, which is easily judged, rather than as two interlaced sine waves, which are not so easily judged. The second feature modulates the level of the two tones at a pseudo voice rate, permitting operation at voice average power levels while viewing a two-tone. (Some transmitter power supplies "droop" during single-tone or even two-tone operation, reaching their peak power output only with a voice waveform with its corresponding low average power.) The third feature is crystal control of the frequencies, which permits accurate frequency measurement of a transmitter without the necessity of generating the carrier.

This generator was conceived as a school project almost 20 years ago. Assigned to build a circuit board, we made a two-tone generator that used two twin-tee oscillators, with the high oscillator locked to the third harmonic of the low oscillator with a small capacitor. When the output of the low oscillator, through a phase shift network to compensate for phase shift in the transmitter, was fed to the X input of a scope and the transmitter output was fed to the normal Y input, a "bow tie" pattern resulted, making judging distortion easy. But the tones had significant distortion and the third harmonic lock was unstable. Over the years I tried to improve it by various means, including the use of function generator ICs. This design, which offers maximal features but minimal complexity, is the final product.

operation

For normal operation the output is connected to the microphone input or phone patch input of the transmitter (see **fig.1**). The oscilloscope is usually connected to the RF output via an X10 probe; the RF output is then connected to a 50-ohm dummy load. Turn the transmitter and the test generator on and adjust the balance and level controls for minimum valleys and peak output as shown in **fig. 2A**. The switches should be in the "normal" and "600 Hz" positions. The scope output may be connected to the "external trigger" or "channel B" input to the scope and used to synchronize the display. Varying the phase control will shift the display horizontally.

For a "bow tie" pattern, shift the scope output of the generator to the X input of the scope. Adjust the X channel gain for full horizontal display. With the transmitter on, a pattern similar to **fig. 3** should occur. Rotate the pattern with the phase control to produce a "bow tie" pattern similar to the one shown in **fig. 2B**. (**Figure 2B** was taken with a commercially built Amateur transceiver and appears to be perfect, yet slight variations in linearity are clearly visible in the "bow tie" pattern of **fig. 2B**.) **Figure 4A** shows a pattern produced by a transmitter with severe crossover distortion caused by low screen grid voltage and **fig. 4B** depicts the same with a "bow tie" display. **Figures**

By Bill McLagan, YB9ATA/WA7AQN, c/o MAF, Box 239, Sentani, Irian Jaya, Indonesia 99000

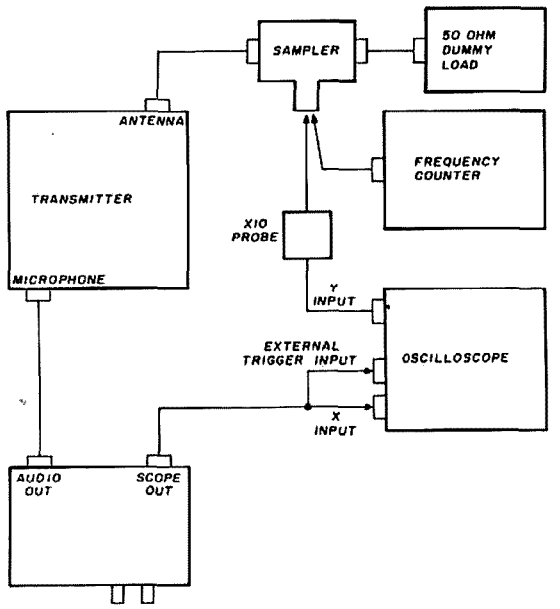


fig. 1. Test connections, two-tone test generator to transmitter.

table 1. Measured power output of a transceiver under test.

Single tone	120 watts
Two-tone	60 watts
Two-tone, pseudo voice	15 watts
Voice ("e" as in "hello")	15 watts
Voice, as above, 20 dB compression	50 watts

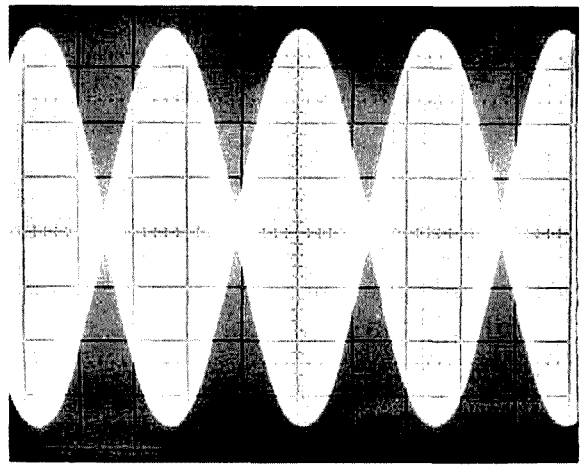


fig. 2A. Output of a commercial transceiver, standard two-tone pattern.

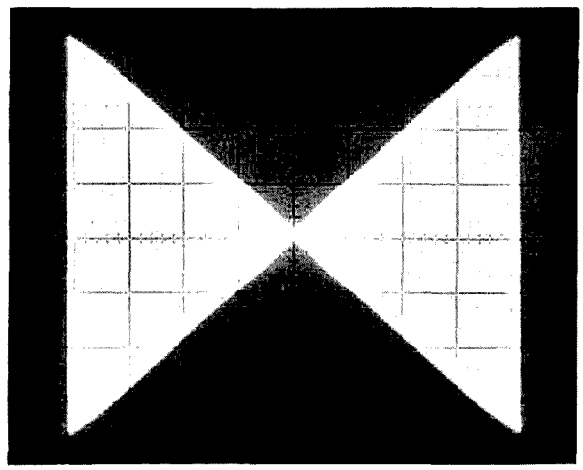


fig. 2B. Output of a commercial transceiver, bow tie pattern.

5A, 5B, 6A, and 6B show two different kinds of distortion and the resulting patterns.

Figure 7 shows a display using the "modulated" feature to reduce the average power to that encountered during modulation. The synch switch is changed to 75 Hz to synchronize the display. Relative power levels for each mode as measured on a commercially built Amateur transceiver are shown in **table 1.**

For precise frequency measurement, a single tone is fed into the transmitter. The audio frequency is added to (for LSB) or subtracted from (for USB) the counter reading to arrive at the correct carrier frequency.

theory, digital section

U1, a CD4060, contains an oscillator and a 14-stage binary divider. Y1 can be either a 3.579545 TV crystal (producing output frequencies of 582 and 1746 Hz) or a 3.6864 crystal (producing output frequencies of 600 and 1800 Hz). The tenth binary stage puts out a 3600 Hz square wave, which is fed to U3, a CD4017 decimal counter set up to reset on the third count, resulting in a 1200-Hz asymmetrical square wave.

Both the 3600 Hz signal and the 1200 Hz signal are divided by two by U2 to produce 1800-Hz and 600-Hz symmetrical square waves. The 1800 Hz signal is locked to either the leading or trailing edge of the 600 Hz wave via differentiator C17 and R4, which triggers the reset input of the 1800 Hz divide-by-two section. S4 permits triggering from either the Q or $\overline{Q}$ output of the 600-Hz divide-by-two. This insures that the generator will always start up with the same phase relationship and permits changing the phase 180 degrees to compensate for the less-than-360 degree phase shift available from the "fine" phase control. U7 further divides down the signal to provide 75 Hz

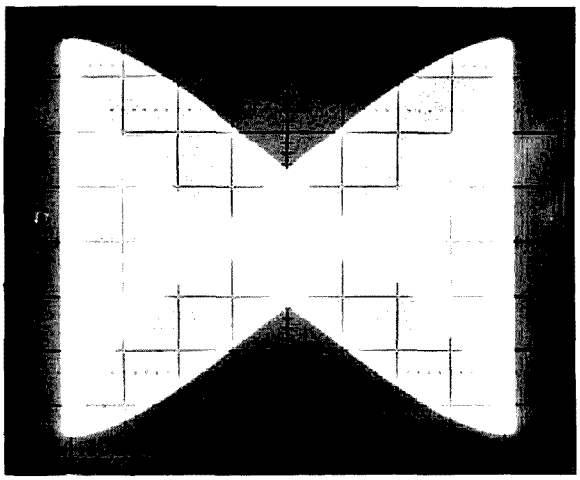


fig. 3. Bow-tie pattern before adjustment of phase control.

for the modulator and 0.146 Hz for the LED "on" indicator. The LED flash provides a visible indication of the operation, yet avoids the 10- to 20-mA drain of a steady lamp.

theory, analog section

R5, C5, R6, and C6 provide two poles of lowpass RC filtering to reproduce a near sine wave of 1800 Hz. I've found that when converting a fixed frequency square wave to sine that multiple RC lowpass sections, operated well beyond cutoff and followed by an op amp to recover gain, is simple and effective. In this case, very low distortion is needed so U4B is configured as a multiple-feedback bandpass filter. U4A is set up similarly, but for the 600 Hz channel. The series resistors in the RC sections may have to be changed slightly to adjust for equal gain through each channel and/or set the output of the bandpass filters

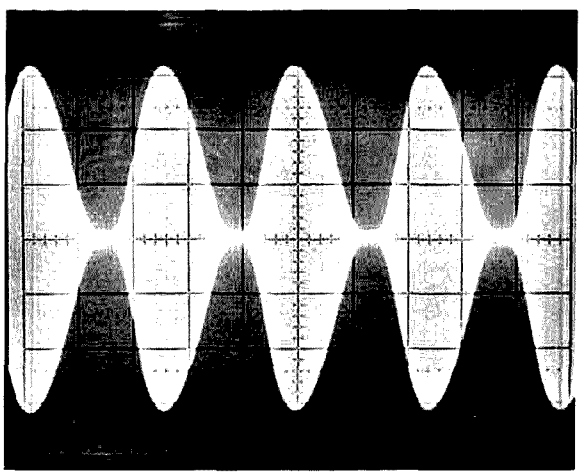


fig. 4A. Severe crossover distortion, standard position.

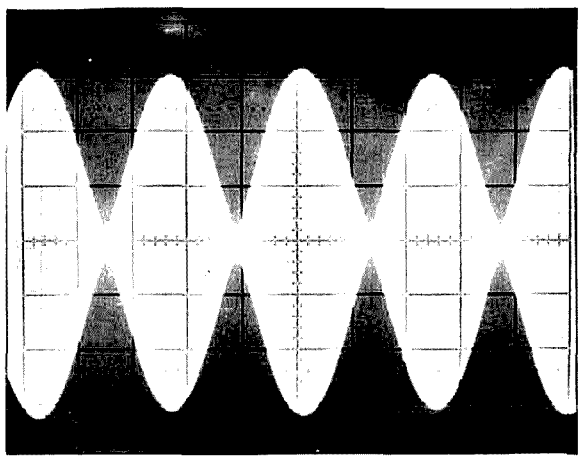


fig. 5A. Minor distortion, higher levels emphasized, standard pattern.

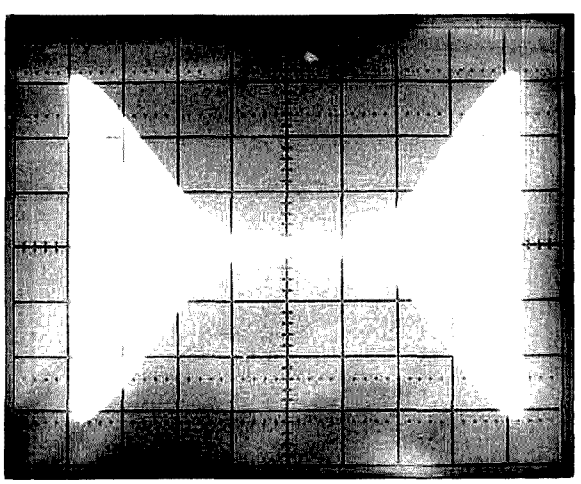


fig. 4B. Severe crossover distortion, bow-tie pattern.

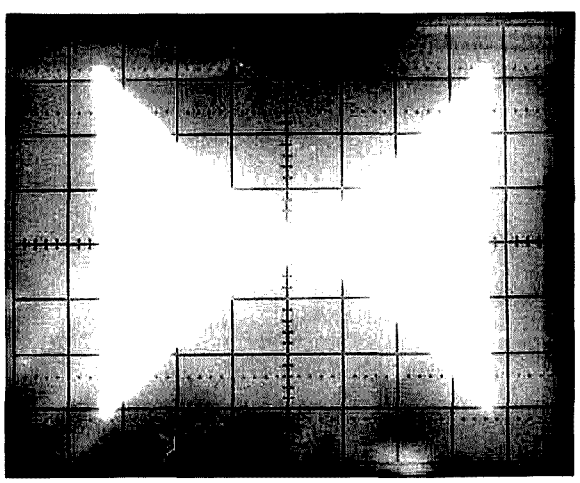


fig. 5B. resulting bow-tie pattern.

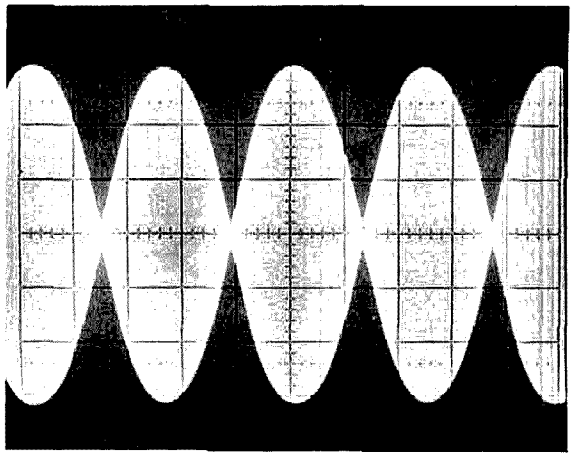


fig. 6A. Minor distortion, higher levels deemphasized, standard pattern.

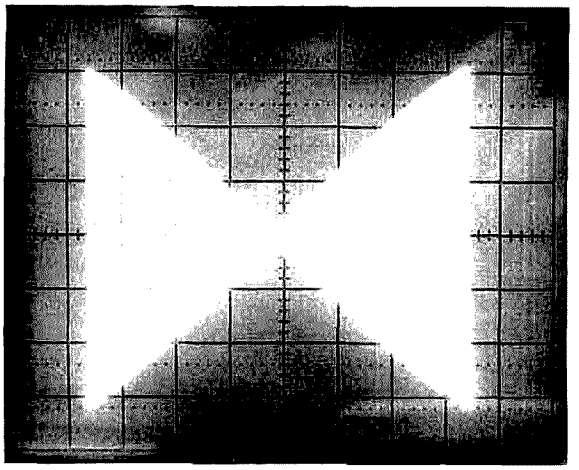


fig. 6B. Resulting bow-tie pattern.

to approximately 1 volt RMS. Both signals are mixed together at the balance potentiometer and go to the the transmitter through a voltage follower, U4C, and the level control. Up to this point, the circuit is similar to any conventional two-tone generator.

To obtain the "bow tie" pattern, the low tone must be sent to the X channel through a phase shifter to compensate for phase shift in the transmitter. U4D inverts the 600 Hz tone and feeds one side of phase shift network C13 and potentiometer R20. Nearly 360 degrees of phase shift is possible. Voltage follower U5B presents a high impedance load to the output of the phase shift network and drives the scope isolation transformer, T1. Note that the secondary of T1 is kept floating from the chassis. Without DC isolation, severe problems with ground loops often occur. Care must also be taken with oscilloscopes that have a simple X input because they can be easily overdriven by the several volts available at the scope output. A 500-ohm to 500-ohm transformer could be used as well.

To provide the modulated function, the 75 Hz square wave is converted to a near sine wave by three poles of RC filtering followed by gain recovery stage U5A. The 75 Hz signal enters the AGC input of U6, varying the output at a 75 Hz rate.

construction

I built the prototype using perf board, IC sockets, and point-to-point wiring using 30 gauge wire. Later I made a circuit board. Either way works fine. In the unit shown (see photo 1), I used some exotic dual-section pots with push-pull switches from my junk box, but normal pots will work just as well. I also added an extra 9-pin plug to allow the use of individual adapter cables to various transceiver microphone inputs and a switch to key the transmitters on. The case, battery holder, and switches came from Radio

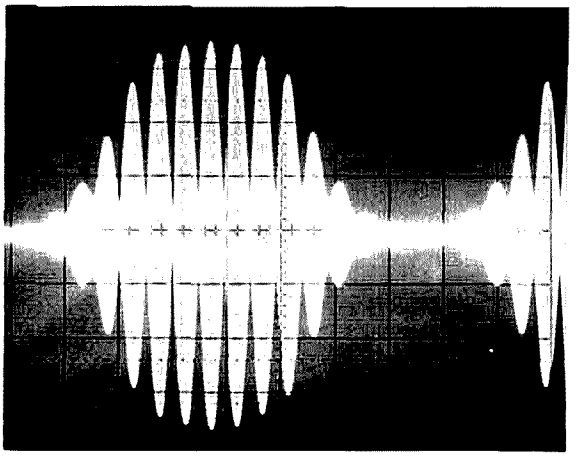


fig. 7. Pseudo voice pattern using "Modulated" functions.

Shack. Mylar capactors should be used in the filter sections and the phase shift network. The 0.1 μF capacitor shown in the power supply line should be placed directly across pin 16 and 8 of U1.

alternatives

One alternative that might be tried is using a 32 kHz watch crystal. In this case, R1 should be changed to 10 Megohms and R2 to 680 or 750K. Only four stages of binary division would be required to produce 4096 Hz at the input of U2A, resulting in input tones of 2048 Hz and 682.66 Hz. The TL084 and LM353 ICs were used because they produce very little distortion. An LM324 and an LM348 could be used if resistors were added from the output of each amp to either ground or V_{cc} to cause the IC to always supply current in one direction. The crossover glitch is very visible on a sine wave. Don't omit the load resistors on the transformers; if you do, significant distortion will result.

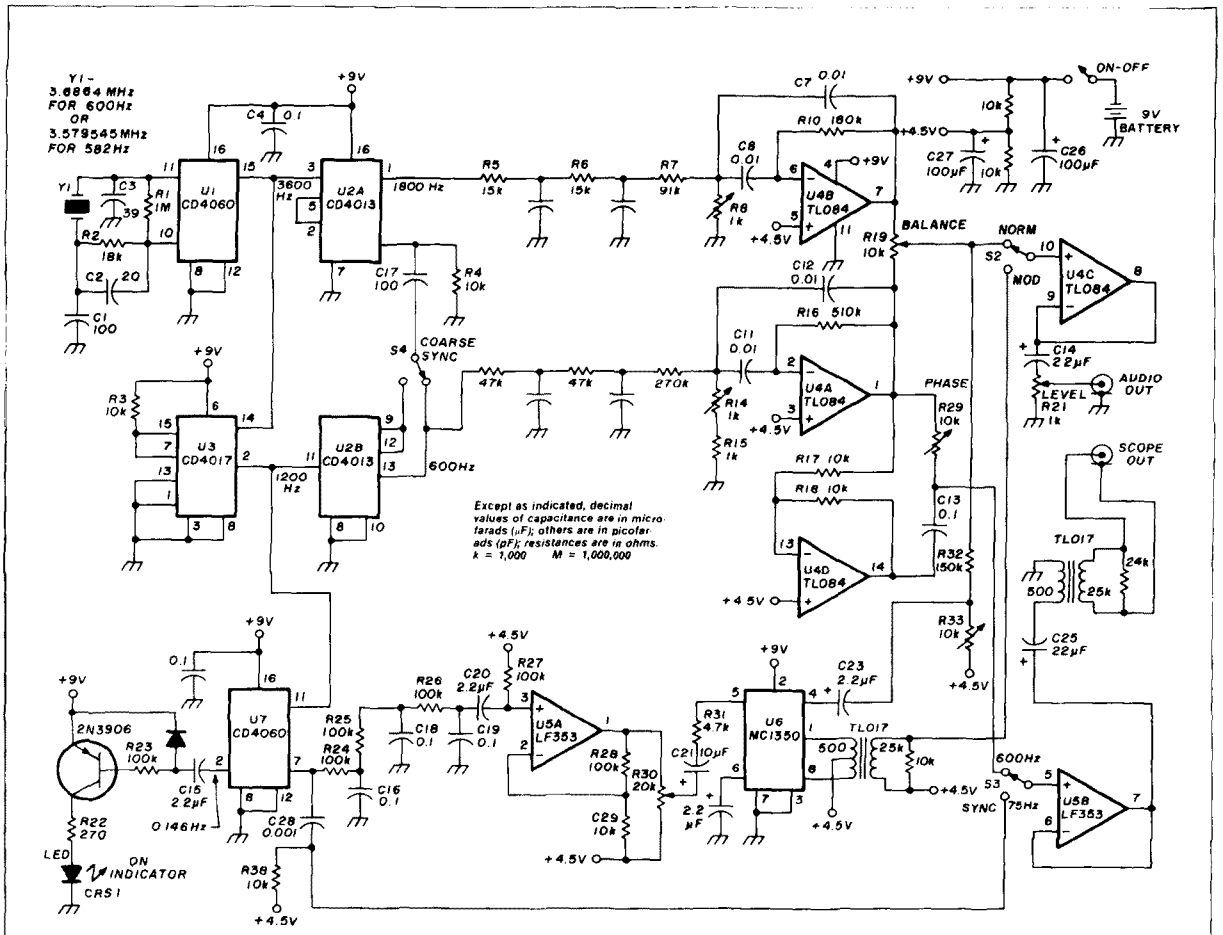


fig. 8. Schematic of two-tone test generator.

If you want to operate the two-tone test generator from an external AC or DC supply, or to ground the case to some external test equipment, an isolation transformer must be added to the audio output. In this case, T1 would not be needed, and the scope could be driven directly from voltage follower U5B. I tried this in one case and it worked fine. Again, don't forget to to terminate the transformer correctly! The power supply can be anything from 8 to 14 volts but must have very low ripple. I would recommend a three-terminal regulator in this case.

testing the circuit

Plug in the digital ICs. Use some form of current limiting in the power supply — to begin with, a 10-ohm resistor in series with a battery or a current limiting power supply. Turn the unit on and determine that normal current is flowing. Check for 3600 Hz, 1800 Hz, 1200 Hz, 600 Hz, 75 Hz, and 0.146 Hz at the places shown. If the lamp begins to flash every 5 seconds or so, then you can be sure most of it is working.

Plug in U4 and U5. Place a scope probe at the output of the 1800 Hz sine wave generator, U4B, and adjust R8 for maximum. Approximately 2 volts peak-to-peak should be present at this point. If the level is too high, causing distortion, and everything else checks out, increase the values of R5 and R6. In the same way, check out the 600 Hz sine wave generator. Move the scope to the output connector. Look for about 2 volts peak-to-peak of combined 600 Hz and 1800 Hz as the balance control is varied (with level at maximum) and for normal operation of the level control. Check at the output of U5A for a 75 Hz sine wave.

Put the scope on the output again, adjusting for 1800 Hz only. Put the S2 in the "mod" position and adjust R33 and R30 to produce approximately 100 percent modulation of the 1800 Hz tone with a peak level equal to the level in the "normal" position. Some interaction is present between these two controls, so several adjustments may be necessary. Turn the balance control for only a 600 Hz tone and check for approximately the same peak voltage in the "mod"

and "normal" positions. The only glitch I've noticed in building several units is a tendency to oscillate when the level control is set at maximum. This occurred on a unit that used ribbon cable to connect the controls to the PC board, and since the level control is rarely at maximum in operation, I decided to ignore it.

conclusion

A two-tone generator with a "bow tie" pattern makes it easy to check linearity in an SSB transmitter. In fact, I found I had to be careful because I ended up trying to correct linearity problems that were normal to the equipment! But on the whole, it has made SSB servicing much easier and is certainly less expensive than a spectrum analyzer.

ham radio

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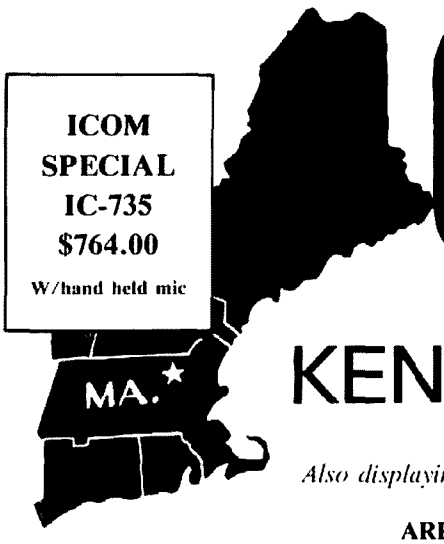
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troubleshooting II: solid state circuits

Last month I discussed methods for isolating the dead stage in a radio receiver. These methods — signal injection and signal tracing — are capable of showing us which stage is bad, and from there we can do further troubleshooting to isolate the faulty component.

In this installment, we'll look at another method that I learned more than two decades ago when solid-state car radios first appeared on the market. This method uses a DC voltmeter to isolate the faulty stage. It's not foolproof, but is well suited to many applications — especially when combined with other methods.

First, let's look at **fig. 1**. Here we have NPN and PNP transistors. In most amplifier circuits the base-emitter voltage will be 0.2 to 0.3 volts for normal germanium transistors (used in older equipment) and 0.6 to 0.7 volts for silicon transistors. In a PNP transistor the base is more negative than the emitter, and in an NPN transistor the base is more positive than the emitter.

Further, the collector will be more positive than either base or emitter in NPN transistors,* and more negative in PNP transistors. Keep in mind that the term "more negative" can mean "less positive" in some cases. I know that's confusing, so let's try to clean it up. Consider **fig. 2**. Here we have a cascade chain of three stages in a radio receiver. Each stage consists of a PNP transistor that's powered from a positive DC power supply. The collectors of the transistors are near ground potential, while the emitters and bases are closer to the +10.5 volt "B+" line. If you measure the collector voltage with respect to ground, you'll find it very slightly positive, while the emitters are at a much higher potential. Thus, the collector's being less positive "causes it to perform as if it were negative."

Voltmeter "A" will measure the potential between the points being measured (in **fig. 1**, an emitter) and ground. If the voltage is near normal at each stage, then we can assume that there are no massive short circuits — but we can't get a hint of whether the stage is working properly.

Voltmeter "B" is connected with its positive electrode on the B+ distribution line, and the other used to probe the emitters of the stages. The B+ line can often be located from the service manual, but in some cases you'll have to fake it. Look for the electrolytic "filter capacitor" used to decouple the B+ line (C1 in **fig. 2**). This filter capacitor will denote the proper line (unless you accidentally selected C2!), and usually has enough of a tab to allow connection of the voltmeter probe.

The voltage drop across the emitter resistor of each stage indicates the current conduction of the stage. If the service manual does not give the normal voltage drop, then calculate it from the emitter potential printed on the schematic and the B+ voltage. In the case of the IF amplifier of **fig. 2**, for example, the emitter voltage is 9.1 volts, while the B+ voltage is 10.5 volts. The normal conduction of this stage will be indicated by a voltage drop across the resistor of 10.5 − 9.1, or 1.4 volts. Any radical departure from this value indicates a problem. For example, a shorted transistor would send that conduction voltage (across R3) up to near 10 volts, while

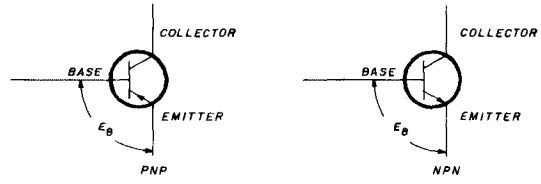


fig. 1. Specific voltage drops across base-emitter and collector-emitter junctions help determine if the transistor is functioning.

*If the transistor stage is in saturation, the collector voltage will drop *below* the base voltage — i.e., $V_{CE_{SAT}}$ and $V_{BE_{SAT}}$

a leaky transistor would place the voltage somewhat lower but still larger than 1.4 volts. Similarly, an open emitter (or other condition that cuts off the stage) will reduce the voltage across the emitter resistor to nearly zero.

We can isolate the defective stage by looking at each emitter voltage in turn. Most often, the defective stage will show up from an anomally in the emitter conduction voltages.

Radios with NPN transistors in the stages are similarly treated. **Figure 3** shows a radio with NPN transistors powered by a positive-to-ground DC power supply. The collectors of these transistors will be close to the B+ potential, while the emitter and base voltages will be a lot lower. The values shown in **fig. 3** are typical, but they are not to be held as absolute (there are a lot of design choices that could alter the values, so buy and consult the service manual for your rig).

As in the case of the PNP stages, the NPN emitter conduction voltage denotes stage activity. In this type of circuit, however, the reference is ground instead of the B+ line. The principle is the same, however. We can check each conduction voltage in its turn and determine whether any of them are off. Fortunately, there are fewer calculations to make in this type of circuit. The emitter voltage on the schematic is the conduction voltage.

Both of the methods discussed above assume that there is a positive with respect to ground DC power supply. This arrangement is used in all modern American automobiles, so will be found typically in mobile rigs. In home rigs the DC power supply is often the same positive with respect to ground as shown here, but that is not necessarily so. It might be the case that the power supply is negative with respect to ground, or that there are two power supplies (one positive to ground and one negative to ground). The principle in these cases are the same, only the point of voltage reference will change.

Variable frequency oscillators, whether in a receiver or transmitter VFO, will behave a little differently. In most cases, the DC conduction of the oscillator transistor varies with frequency setting. Typically, the voltage will be higher at the low end of the band, and lower on the high end of the band (with a smooth transition as the dial is tuned). A sudden discontinuity in this transition might indicate a sud-

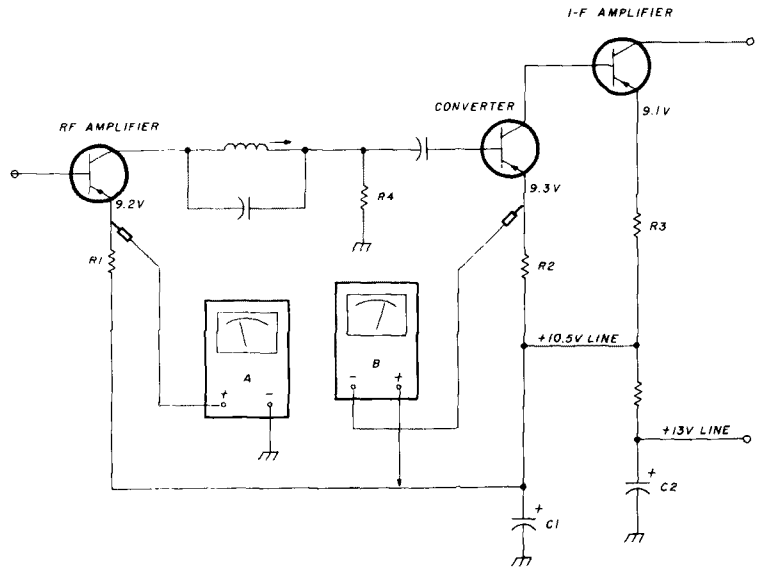


fig. 2. Voltage drops across the emitter resistors determines if the PNP transistors are operating properly.

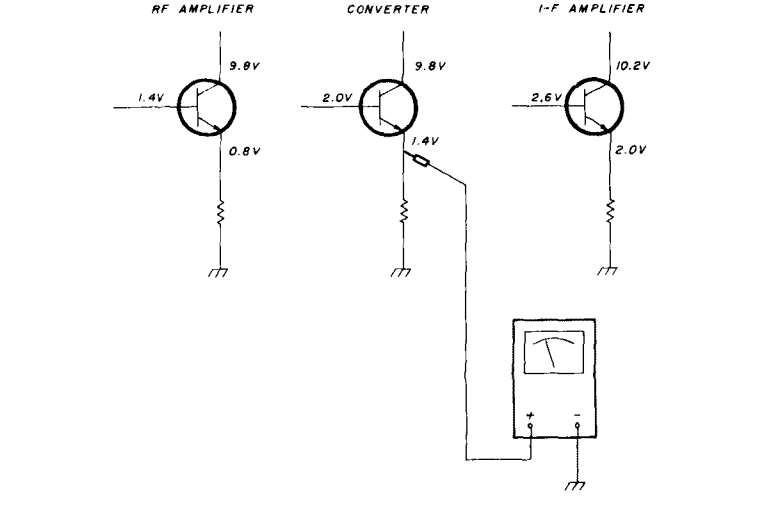


fig. 3. With NPN transistors, emitter resistor voltage drops are usually determined just by measuring the emitter voltages (with respect to ground).

den cessation of oscillation (or a parasitic developing) at that point on the dial.

Crystal oscillators behave in a similar manner. If the crystal selector switch is changed, then the voltage can be expected to change also. If the crystal is removed, then the voltage will usually change radically.

By the way, don't remove crystals willy-nilly. In some cases, especially when the next stage is a grid-leak biased vacuum tube power amplifier, damage might occur to the equipment.

an ambiguous case

There are three ways to stage isolate a dead radio receiver. In part I of this two-part series, we reviewed signal injection and signal tracing and in this article we discuss using a DC voltmeter to find the defective solid-state stage. But some situations are not as clear.

Consider **fig. 4**. This circuit is a simplified audio section, with only the preamplifier shown in detail. The combination of AF preamplifier and output power amplifier may be capacitor coupled (as shown here), direct coupled or in the form of a single integrated circuit housing both stages. The symptom is weak reception. All stations are heard, but at very low volume. If you read last month's article, you'll recall that this symptom generally points to the output IF amplifier, detector or audio stages; but in this case, it's the audio stage that contains the fault.

I've seen a lot of cases where the DC voltages on Q1 were normal and there seemed to be adequate signal passing through the stage regardless of whether signal injection or tracing was used. The ambiguity will show up except in the very special case where the sensitivities and impedances of the signal generator and signal tracer are matched to the particular equipment being serviced (some "radio analyst" instruments used by commercial servicers fall into this category).

In the event that this ambiguity shows up, jump on capacitor C2. This capacitor is used to place the emitter of Q2 at or near ground potential for AC while retaining the DC bias on the emitter. If this capacitor opens up, the gain of the stage will go down one heckuva lot without (usually) affecting the DC voltages. Turn the rig off, solder tack a replacement capacitor across C2 (the exact value isn't critical — anything ±200 percent of the correct value will do for the test). Turn the rig on. If the output volume comes up a lot, then mark C2 for replacement.

I don't know why it is, but this problem seems to keep popping up. I've seen it in varying degrees in car radios, home radios, Amateur radios, CB radios and commercial two-way radios. It also turns up in vacuum tube equipment.

Old Gear Restorer's Note: If you restore an old Heath DX-60B (or other older transmitter) and find the RF output intermittent as the rig is keyed, then look to the plate circuit of the crystal oscillator. This circuit is a modified Miller oscillator, and has a slug-adjusted inductor in the plate circuit. When the rig is stored, unused for several years, the core of this coil will change characteristics, and the oscillator will refuse to oscillate reliably. I scratched my head over two of these rigs last year and finally stumbled on the answer. While keying the rig, carefully adjust the slug until reliable operation occurs (as indicated by a lightbulb dummy load or RF wattmeter), then give it another turn or so.

Incidentally, similar circuits are used in a lot of older ham gear, both vacuum tube and solid state, HF and VHF. Anytime you have a piece of equipment with an oscillator of this type that seems to operate intermittently or unreliably, and everything else seems to check out, then consider adjusting the coil.

Normally I don't recommend "diddlestick" troubleshooting — to the extent of hollering "NEVER!" when someone suggests it. One of the few possible exceptions might be in equipment that has been unused for many years, and contains ferrite core coils in oscillator circuits. I still maintain that sudden failures in equipment that has been working should be regarded as a component failure, not as an alignment problem so in that case, leave the darned diddlestick in the toolbox!

ham radio

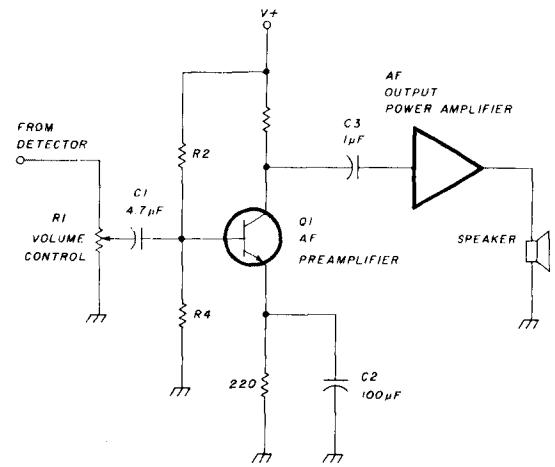


fig. 4. Loss of gain in audio stages can sometimes be attributed to an open emitter bypass capacitor.

the offset drooper
an improved ground plane

Reduce antenna effect and still achieve a 50-ohm feedpoint resistance

The original "ground plane" omnidirectional antenna was developed in the late 1930s jointly by Dr. George H. Brown, J. Epstein and R. F. Lewis, W2EBS, all of RCA Laboratories. It consists of the long-familiar configuration of a vertical quarter wave "spike" working against four resonant radials at 90 degrees to the mast. In the original version, patented in 1941, the feedpoint impedance is matched to the coaxial feedline by means of a quarter wave coaxial "Q" section. A typical Amateur adaptation is shown in **fig. 1**. In 1942 Dr. Brown patented, solely, the version shown in **fig. 2**, which has several advantages over that of **fig. 1**.

These basic "ground plane" antennas exhibited less antenna effect (surface current) on the mast and feedline than did other omnidirectional VHF antennas widely used at the time. However, the four radials at 90 degrees to the mast are somewhat "transparent" and definitely resonant. For this reason they may be considered more a part of the antenna than a virtual ground plane for the "spike." Semi-infinite ground plane characteristics would allow little radiation below zero degrees elevation. Such a pattern might be unsuited to a mountain top location or a swaying free-standing mast.

Because of the direction of current flow in the radials, the resultant radiation from the individual spikes results in good cancellation, and there is little net radiation from them. The fact that one may touch the tip of one radial with little effect on VSWR doesn't mean that the radials do a good job of simulating a large, flat conductive sheet. Because of the quarter wavelength of the radials, pinching the tip of one elevates the impedance at the opposite end and effectively isolates it. The remaining three radials simply take over, with only moderate detuning of the antenna.

The result of all this is as follows: the performance of a "90 degree" ground plane represents an improvement over several previously popular base station antennas. However, while orienting the resonant radials at 90 degrees does reduce inductive coupling to the coaxial line and mast, it doesn't effectively eliminate it. The coupling is sufficient that the resulting antenna effect may be found undesirable for some applications.

Long popular among hams for VHF and upper HF is the "droopong" ground plane, a simplified, less elegant, low-cost descendant of the Brown ground plane. Simply bending down the horizontal radials to about 45 degrees raises the radiation resistance to 50-52 ohms. This permits direct connection of the coax without use of a matching device. A more appropriate description might be "skeleton skirt dipole," because the resonant drooping radials don't do a very good job of serving as a virtual ground plane. Instead, they exhibit more inductive coupling to the feedline than do radials at 90 degrees.

Nearly 40 years ago, in the *Antenna Manual*, I pointed out that while the drooping ground plane or "drooper" is simple and works well, the configuration aggravates antenna effect. For the benefit of those not familiar with the term, *antenna effect* (transmit case) can be described briefly as follows:

Antenna effect on a two-wire line: refers to line radiation as a result of "common mode" current. Part of the power (energy) fed to the feedline travels on it as though the two wires were tied together and were working against ground. Usually it results from load imbalance and/or excessive coupling from one side of a balanced antenna to the line. Common mode current adds in one wire and subtracts in the other. This in-phase component of the total power fed to the line acts as though the two wires were in parallel. So it gets radiated.

Antenna effect on a coaxial line refers to radiation from a coaxial feedline as a result of current flowing on the outside of the outer conductor. The current often is shared with the surface of a metal mast supporting a bottom-fed vertical antenna. Usually it results from either or both of the following:

- The outer conductor of the coax is directly connected to a point on the antenna not precisely at

By Woodrow Smith, W6BCX, P.O. Box 2898, Anaheim, CA 92804

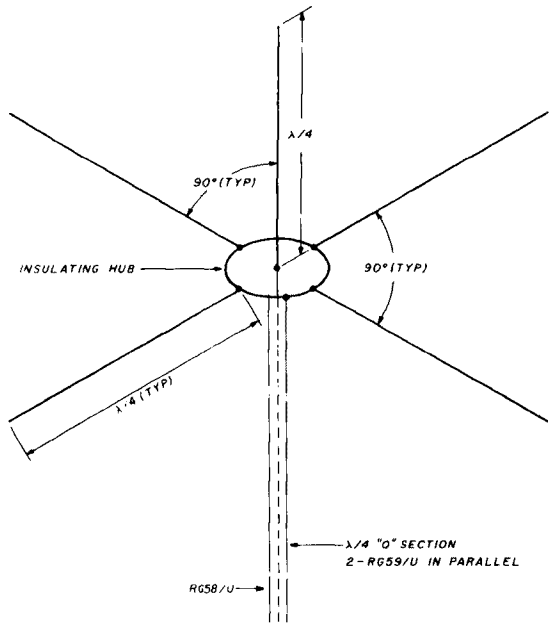


fig. 1. Typical Amateur adaption of the classic (original) Brown ground plane antenna, suitable for upper HF use on a flat roof, or for upper VHF use on a mast.

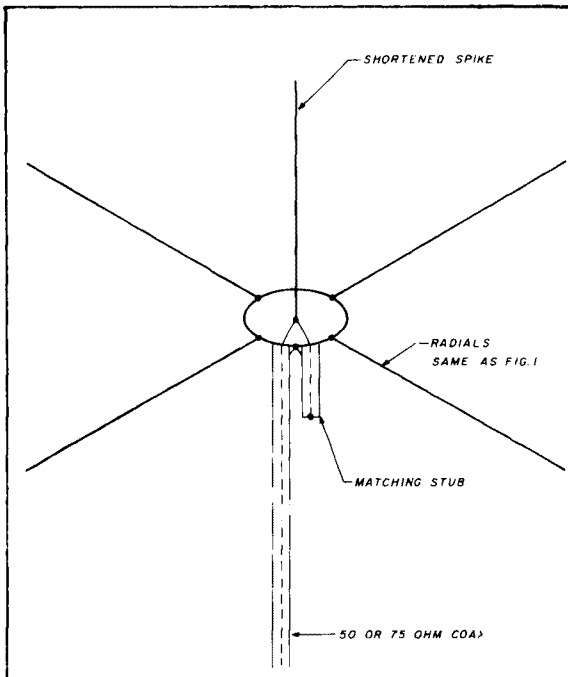


fig. 2. A later version by Brown eliminates the Q section and allows easy tweeking for a perfect match at a spot frequency. Useful where the antenna must be side-mounted on a shared mast of large cross section, or is other wise subjected to unpredictable detuning.

ground potential. As a result, current flows not just to the antenna, but is encouraged to flow back down the *outside* of the coax shield to a virtual ground as well. The "ground" can be a cabinet at VHF or house wiring at HF, for example.

• Excess inductive coupling exists from one half of the antenna to the outside of the coax or metal supporting mast (or both). The radials of a drooping ground plane are a case in point. If the coax is enclosed by a metal mast, spurious current flows on the surface of the mast. If the coax is draped along the outside of a metal mast, current flows on the surfaces of both and *both* radiate. To complicate things, the mast can have its own virtual ground.

is antenna effect all that devastating?

At just what point antenna effect becomes serious enough to be concerned about is debatable. The amount of current that can be tolerated on the outside of a coax line or mast or both depends to a great extent upon the following:

On transmit, let's first consider a well elevated, vertically polarized, omnidirectional VHF antenna with a feedline many wavelengths long. What is the result of antenna effect? Normally little of the power (energy) radiated with vertical polarization from such a long feedline (and the mast) will be directed at the horizon. For this reason, little of the spurious radiation will either add to or subtract from the energy being radiated effectively towards the horizon by the antenna proper.

So at VHF the end result of antenna effect on *transmit* is primarily a waste of power. But even if as much as 20 percent of the total radiated power is radiated by the coax, mast or both and thereby wasted, the resulting 1 dB loss is hardly anything to get worked up about. This is especially true when it buys worthwhile simplicity, convenience, or economy. However, if the radiating coax passes close to a TV receiver feedline, TVI may result if the TV coax suffers from poor shielding, or if the TV twinlead feeds a poorly balanced receiver front end.

On *receive*, the situation is different. Consider a coax line running through a localized area of high ambient noise. Antenna effect can cause noise picked up by the outer conductor of the coax to travel up to the antenna proper, then back down the coax to the receiver input just like a desired signal.

baluns vs. resonant isolators

Baluns of the type used with HF dipoles to minimize antenna effect are not suited for use with vertically polarized VHF antennas of the omnidirectional type. Instead, resonant detuning sleeves, cones, and radials are widely used as isolators or decouplers to "cool off" the mast and feedline. To what extent these produce any practical benefit in a particular installation by

reducing antenna effect often is open to question. The use of coiled coax, ferrite beads, or a ferrite sleeve to choke off or dissipate surface current on the coax does not solve the metal mast problem.

Granted, no startling increase in transmitted signal strength will be noted when a resonant isolator of some kind is added to a simple VHF drooping ground plane. But even so, suppose it were possible to achieve a big reduction in antenna effect on the drooper without adding a resonant isolator. Suppose the various resulting advantages (such as they are) could be achieved by simply altering the dimensions and the droop angle.

something for nothing?

It's not only possible; it's *simple*. And there are no additional parts or materials, and without any additional manufacturing, construction, or assembly labor. Here's how: take the case of a conventional drooper that's supported by a metal mast enclosing the coax. Current is induced on the mast as a result of inductive (mutual impedance) coupling to the radials. If the coax exits the hub external to the mast (offset antenna mount), spurious current also appears on the outside of the coax.

This detrimental inductive coupling can be reduced somewhat by reducing the droop of the radials (the angle they make with the horizontal). The remainder can be compensated for by deliberately introducing a critical amount of conductive coupling of opposite phase. This is done by drastically offsetting the feedpoint from the voltage node.

The offset required for good cancellation is accomplished by simply making the radials as much as 30 percent (yes, thirty percent!) longer than the spike. Because precise cancellaton is somewhat frequency-sensitive, the effectiveness will vary a bit over the 2-meter band. However, in spite of the fact that optimized coupling neutralization is less than perfect over the whole band, the practical results obtained are most worthwhile.

Offsetting the feedpoint from the voltage node raises the feedpoint resistance. On the other hand, lessening the droop angle lowers the feedpoint resistance (by lowering the radiation resistance). By proper choice of these two values, it's possible to reduce antenna effect dramatically while at the same time achieving a 50-ohm feedpoint resistance. With an offset represented by a 28-30 percent radial-to-spike length differential, a 27-29 degree droop angle provides both maximum reduction in antenna effect *and* a 50-ohm feedpoint resistance. To some extent the optimum values vary with conductor diameter, mast diameter, and hub geometry. Less important is whether the coax departs the hub inside or outside the tubular mast.

Neither the 1.5 VSWR bandwidth nor the antenna gain is noticeably degraded by drastically offsetting the feedpoint and reducing the droop of the radials. And in case you're skeptical about the horizon gain comparisons, it's true that the spike is slightly shorter on an offset drooper. Likewise, the resultant vertical component or vector of the drooping radials is a little shorter for an offset drooper than for a regular drooper. But careful measurements show this does not affect the gain significantly. Increased current throughout the antenna resulting from the lower radiation resistance (about 35 ohms) compensates. Also, power wasted by line and mast radiation is reduced to insignificance.

Out of curiosity, a check was made to see just what would happen if the feedpoint of a classic 90-degree ground plane were deliberately offset. When the resonant radials were lengthened, an arbitrary 12 percent and the spike length and matching adjustment were re-optimized, antenna effect was virtually eliminated. However, overall performance was no better than that of an offset drooper, which has the advantage of requiring no matching device. Because of the latter, no further work was done with 90 degree ground planes. It is interesting to note that when resonant radials at 90 degrees to the mast were shortened experimentally by no more than a few percent, antenna effect was markedly aggravated.

improvement over what?

A check was made on drooper dimensions given in various handbooks and past magazine articles. Also, two different commercially manufactured 2-meter droopers were purchased and the dimensions measured. The effective radial length ranged from slightly longer than the spike to slightly shorter. All were series-fed at the hub without benefit of a matching device. None employed a resonant decoupler, such as an extra set of radials below the antenna oriented 90 degrees to the mast.

The reduction of antenna effect provided by drastically offsetting the feedpoint of a drooping ground plane depends, among other things, upon how bad the antenna effect is to begin with. This varies somewhat with location of the voltage node on the regular drooper used for comparison. For instance, antenna effect definitely will be worse if the radials of a regular drooper are significantly shorter than the spike, as was the case with one of the name brand antennas tested. This makes an offset drooper appear just that much better by comparison.

a spurious top-fed Marconi is the culprit

The degree of improvement will also vary with the length of the coax. This is explained as follows: the

outside of the coax shield will do its best to act like a "harmonic Marconi" fed at the top instead of the bottom. Just how effective (and therefore how objectionable) this is will depend to a great extent upon the electrical length of the coax shield in wavelengths from "virtual ground" to the point of attachment to the radials.

Unfortunately, with 40 feet of coax, QSY from 144 to 148 MHz can change the electrical length of the shield substantially. This in turn affects the feedpoint impedance of the spurious "upside down harmonic Marconi." The variation can affect the amount of line radiation by 10 dB or more in the case of a conventional drooper. Things get even more involved when comparing an offset drooper to a regular drooper for antenna effect simply by swapping them at the end of the same feedline. The radials of a 2-meter offset drooper are longer than a quarter wave and their impedance to ground at their feedpoint therefore is affected. This was taken into account when designing the antenna and making comparison measurements.

the test set-up

For reference, a 2-meter drooper with optimized offset feed was constructed with four radials 30 percent longer than the spike and having a droop angle of 28 degrees. For starters for those who might like to experiment, the 1/8 inch diameter spike measured 18 3/8 inches and the 3/32 inch diameter radials 24 inches, all of brass welding rod. Size and configuration of the hub will affect the lengths, especially that of the spike. Note that while the spike of an offset drooper is only slightly shorter than normal, the radials are much longer. The result is that the overall length of the spike plus a radial is nearly 10 percent longer than for a conventional drooper. This involves the integral coupling neutralization and impedance transformation process, and a rigorous explanation is not within the scope of this article.

The offset drooper *reference* antenna just described was compared to four different conventional 2-meter droopers for antenna effect, 1.5:1 VSWR bandwidth, and field strength at zero elevation angle. One of the conventional droopers was constructed to dimensions specified in a magazine article. Two were dissimilar name brand units. None employed a detuning sleeve, cone, or extra set of radials. Measurements were taken near 144, 146, and 148 MHz with four feedline lengths differing by 1/8 wavelength.

Tests were first run with the coax leaving the hub contained within a 3/4 inch O.D. mast for the first 5 feet. The tests then were repeated with the hubs offset from the top of the 5 foot upper mast section. With the hubs offset, the coax was brought down snugly against the outside of the mast for its entire length. Next, the hub was mounted concentrically atop a 12-foot section of aluminum tubing strapped to a steel vent pipe, with the coax brought down inside the tubing. While these changes did cause the readings to change, the overall improvement exhibited by the offset drooper did not change significantly. The coax employed was RG/8X-8M 0.25 inches OD, 52 ohm.

test results

A spurious RF current sniffer was improvised to quantify the amount of improvement exhibited by the offset drooper. The sniffer was provided with a plastic spacing fixture that allowed choice of two spacings in order to increase the useful range. It was checked for directional effect (by reversing it) and the directivity was found to be negligible. Relative calibration in dB was accomplished by simply varying the measured power fed to a leaky dummy load which was space-coupled to the sniffer.

The reduction in antenna effect when using the offset drooper exceeded 11 dB for three of the comparison antennas tested. The improvement obtained over the fourth regular drooper (the one with the longest radials) measured 10 dB. The greatest improvement was observed when the offset drooper was compared to the regular drooper having the shortest radials (about 5 percent shorter than the spike). The improvement figures reflect those obtained or exceeded with the worst case combination of frequency, line length, and mast and feedline configuration.

This article is not intended to show the reader how to build something exactly like the author's, but instead to explain a simple method of improving the performance of the venerable drooping ground plane. Just make the spike a little shorter, the radials a lot longer and bend the radials up a bit. It is applicable to modification of existing antennas as well as to new construction. To approach the maximum possible improvement, all you need is a VSWR meter. Just make the radials 30 percent longer than the spike and droop them 28 degrees (or as close as you can). Optimum length for the spike is usually about 6 percent shorter than for a regular drooper that uses similar hub and element diameter.

If this doesn't result in equal VSWR at the band edges, the spike is too long or too short. Once the radial length and droop are optimized, the center frequency can be fudged about 1 percent just by trimming the spike.

The reader will notice that no VSWR numbers are given for an optimum offset drooper. The reason for this is that a conventional drooper with bad antenna effect can give specious VSWR readings, making any offset drooper comparisons meaningless. The deceptive readings obtained with a standard drooper will change with coax and mast lengths, and the readings

can be considerably better (or worse) than the true antenna VSWR. To a lesser extent this applies also to a classic 90 degree ground plane. But with an optimized offset drooper, readings will change little with line and mast lengths (except as reduced by line attenuation), and can be relied upon. The true VSWR of a properly optimized 2-meter offset drooper will approach 1.0:1 over much of the band, and be found very low at the band edges.

so why not?

In closing, I hope the reader accepts the fact that while a bad case of antenna effect on a VHF drooper isn't necessarily disastrous, minimizing it certainly can't hurt. It's easy to do and all it can do is good — so why not?

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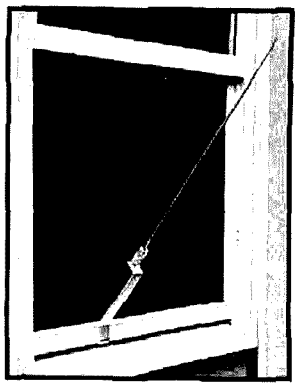
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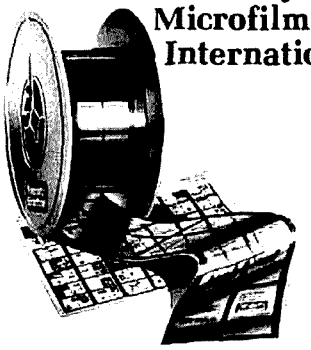
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VHF/UHF WORLD

Joe Reisert W1JR

reflector antennas: part 1

The subject of reflector antennas in general and those of the parabolic dish type in particular generates lots of correspondence and requests for information. I never knew that any single VHF/UHF/ SHF antenna could be so much on everyone's mind — especially since so much material has already been written on the subject. Confusion seems to reign.

Since it began, this column has concentrated on providing solid groundwork in the lower frequencies; limits of time and space have not allowed discussion of more narrowly-defined topics such as specific types of antennas. But with last month's column, we launched a discussion of the microwave bands[1] so this would seem to be an appropriate time to address the subject of reflector antennas since they're so often used at UHF and microwave frequencies.

This discussion will be divided into two parts. Part I — this month's column — addresses the electrical aspects of reflector antenna design, including gain and specific parameters of the parabolic antenna. Part II — next month's column — will cover feed systems and the mechanical aspects of parabolic antenna design. With Parts I and II in hand, you should have all the required material and references necessary to design and build a first-class parabolic antenna.

reflector antenna types

Before discussing the parabolic dish, it should be interesting to review some of the more common types of reflector antennas. They include, but are not limited to, the cylindrical parabola, parabolic dish, corner reflector, backfire, spherical dish, and dual reflector types of antennas.

The *cylindrical parabola* (**fig. 1A**), believed to be the first RF-type of reflector antenna, was used by Heinrich Hertz in 1888! He illuminated a reflector with a spark-gap at about 66 cm (445 MHz) in an attempt to prove the existence of electromagnetic waves, which had been theoretically predicted by James Clerk Maxwell.[2] For best performance, the cylindrical parabola requires a line feed. (Thus was born the field of microwave optics.) An Amateur cylindrical parabola is described in reference 3.

Reflector antennas were soon eclipsed by the search for DX on the HF (high frequency) bands by Marconi and others. Thus it wasn't until the 1930s that serious work on reflector antennas got under way, primarily through the efforts of Grote Reber, ex W9GFZ, in his radioastronomy experiments. Reber used the parabolic dish type of antenna shown in **fig. 1B**. Properly designed, it focuses all the energy at a common point, which greatly simplifies the feed system. Hence it is often referred to as a prime focus-fed antenna.

The corner reflector, as depicted in **fig. 1C**, was invented by Dr. John Kraus, W8JK, in 1940.[4] However, it is limited to gains of up to about 15 to 20 dBi and is quite large relative to the gain delivered.

The *backfire antenna* was invented by Dr. Herman Ehrenspeck in 1958.[5] He discovered that if a Yagi antenna were aimed into a special reflector system and the Yagi structure properly retuned, its gain could be increased by 3 to 5 dB **(fig. 1D)**. However, the backfire antenna never gained popularity because it required a large reflector with a special stepped surface.

Later Dr. Ehrenspeck designed the *short backfire antenna* shown in **fig. 1E**.[6] It's smaller and simpler, but its gain is lower (16 dBi is typical). However, it has gained popularity with the military, especially for portable VHF/UHF satellite operation.

At first glance, the *spherical reflector* looks very much like a parabolic dish. But it has a different reflector profile and is very difficult to feed because the energy is not focused on a common point, as in the parabolic dish. Instead, the energy is present along a line, and a special phased line-source feed is required to properly phase the energy.

Because of its shape, the spherical reflector beam can be partially aimed by simply moving the feed system. This allows the beam to be steered off boresight by up to several times the antenna beamwidth, with little or no performance degradation. Hence the spherical reflector antenna is often used in large structures where the reflector is stationary.

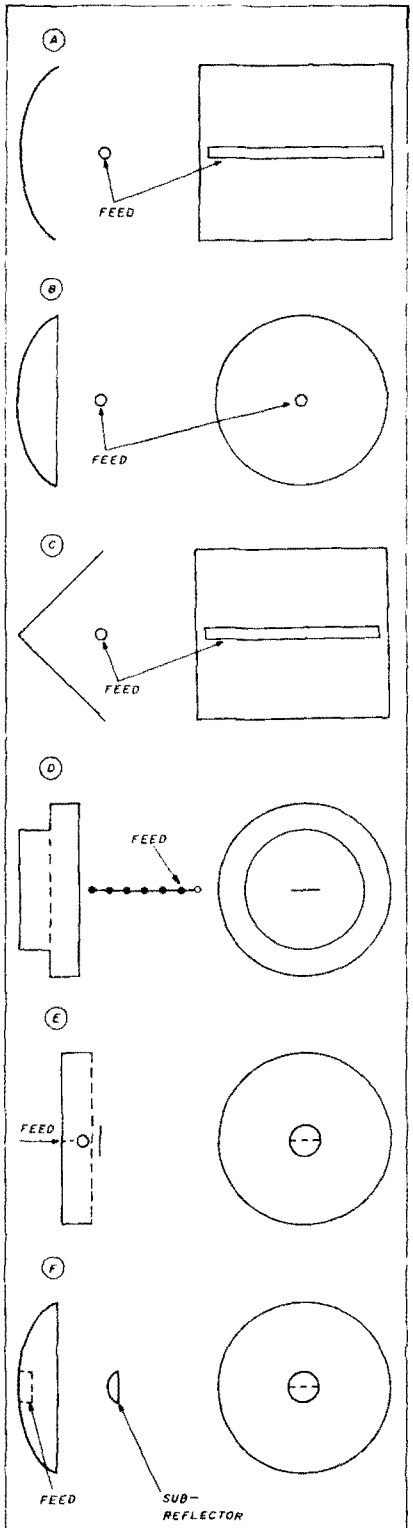


fig. 1. Some common types of reflector antennas. *A*) cylindrical parabola, *B*) parabolic reflector, *C*) corner reflector, *D*) long backfire, *E*) short backfire, and *F*) dual-reflector.

The largest and certainly most famous spherical reflector antenna is at Arecibo, Puerto Rico. Built in the crater of an extinct volcano, it measures 1000 feet (305 meters) in diameter, has a stationary reflector, and uses a special "scalar" feed that can be steered. At 430 MHz, this feed is over 96 feet (28 meters) long![2]

Other antenna systems use a secondary reflector that allows the feed system to be placed at a more convenient point in the structure (see **fig. 1F**). Examples are the *Gregorian*, *Cassegrain*, and *Newtonian* fed-reflector antennas. There are many other variations of the antennas just described.[2,7,8,9] These variations are associated with the different feed system employed.

The implementation of a secondary type of reflector is usually beyond the reach of the typical Amateur and beyond the scope of this column. The improvement obtained with such systems is often at the expense of considerable complexity, which can often be offset by designing a slightly larger antenna right from the start — certainly a typical Amateur approach!

tradeoffs

All types of reflector antenna systems have their advantages and disadvantages. The technical literature on reflector antennas could easily fill a good-sized room! (See references 2, 7, 8, and 9 for starters.) But for the Amateur, the parabolic reflector antenna that uses a prime focus feed system as shown in **fig. 1B** is one of the simplest antennas for obtaining high performance with relatively straightforward construction.

Some of the main advantages of the parabolic dish include its use of a single feed; its ability to operate on other frequency bands by simply exchanging the feed system; its low noise pickup — i.e., its clean radiation pattern. It has reasonable gain for the size of the aperture employed. Its disadvantages include large surface area, which leaves it vulnerable to damage

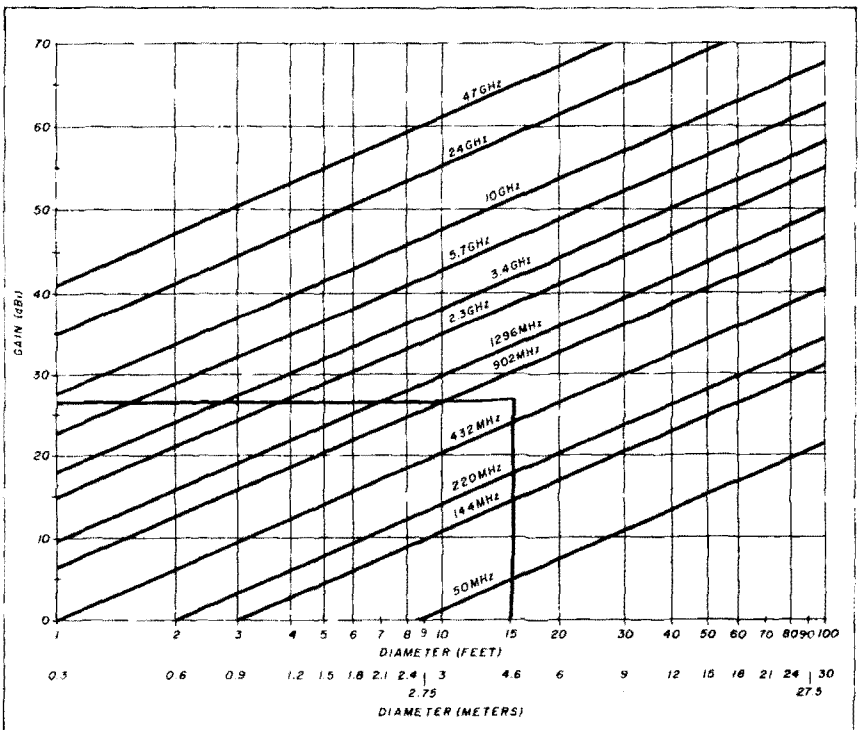


fig. 2. This graph is a plot of gain versus size for typical parabolic dishes with 55 percent efficiency.

from wind or snow; comparatively large physical size (for gain) than a Yagi array offering equivalent gain; and the mechanics involved in designing the reflector and rotation system, especially if it's used on EME. Despite these shortcomings, the parabolic dish, with its single feed system, is a natural for the UHF and especially the Amateur microwave bands.

gain and beamwidth

Generally speaking, the efficiency of a well-designed parabolic dish is in the range of 50 to 60 percent (more on this later). Hence it will have an effective aperture or capture area that is somewhat smaller than its physical area.[10]

Based on a 55-percent efficiency, a typical design standard, the gain of a well-constructed parabolic dish can be easily calculated using **Eqn 1**.

$$G = 10 \log [0.55 \cdot 4\pi A/2] = 10 \log (6.9A/2) \quad (1)$$

where G is gain in dBi and A is the area of the reflector. Both area and wavelength must be in the same unit of measure.

For example, the area of a 20-foot (6.1-meter) diameter dish is about 314 square feet (29.2 square meters). At 432 MHz a wavelength is 2.277 feet (0.694 meters). Therefore, using **Eqn 1**, the gain would be about 26.2 dBi.

You may prefer to work directly in feet/meters and frequency. If so, the following formulas can be derived from **Eqn 1**.

$$G = -52.4 + 20 \log D + 20 \log F \quad (2)$$

where gain is in dBi, D is dish diameter in feet and F is frequency in MHz or

$$G = -42.1 + 20 \log D + 20 \log F \quad (3)$$

where G is in dBi, D is dish diameter in meters and F is in MHz. **Eqns 2** and **3** yield the same gains as **Eqn 1** without having to first calculate the reflector area.

In order to make it easier to rapidly determine gain, I have prepared the graph in **fig. 2**, which includes practical size reflectors and gains of 10 to 70 dBi on most of the popular Amateur VHF/UHF bands. Note that in the example just described, the graph shows the gain to be about 26 dBi, very close to the calculated value!

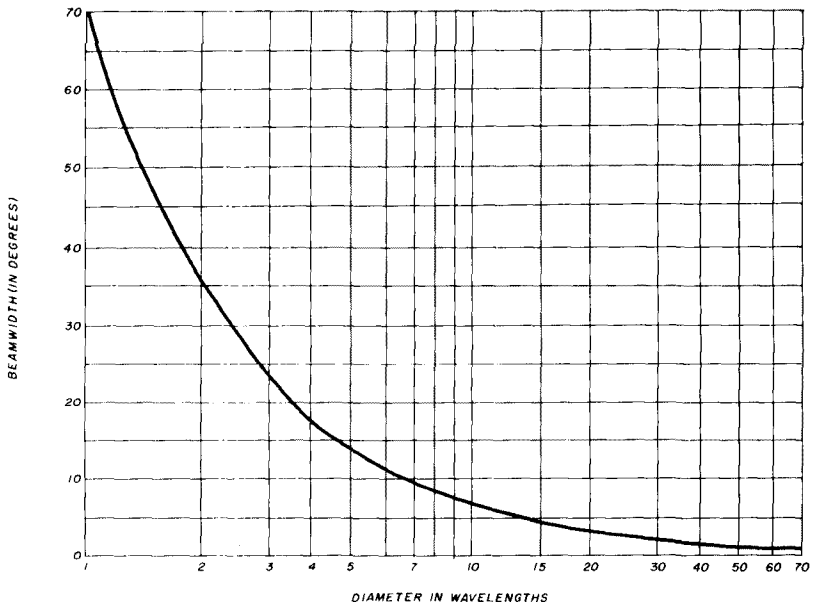


fig. 3. Antenna beamwidth versus parabolic dish diameter in wavelength.

table 1. Typical parabolic curve dimensions for a 20-foot (6.1 meter) diameter parabolic dish with a 0.5 f/d ratio.

distance from center of dish (Y)		depth of dish (X)	
feet	meters	inches	centimeters
1	0.305	0.3	0.76
2	0.61	1.2	3.05
3	0.915	2.7	6.86
4	1.22	4.8	12.19
5	1.52	7.5	19.05
6	1.83	10.8	27.43
7	2.13	14.7	37.34
8	2.44	19.2	48.77
9	2.47	24.3	61.72
10	3.05	30.0	76.2

Finally, it has been pointed out before in this column that as the gain of an antenna increases, the beamwidth decreases.[11] For a parabolic antenna, the beamwidth can easily be determined to a reasonable accuracy (*± 10 percent*) using the following formula:

$$BW \approx 70\lambda/D \quad (4)$$

where BW is in degrees and D is the reflector diameter. Both wavelength and diameter must be in the same unit of measure.

For example, the antenna just described has a diameter of approximately 20 feet (6.1 meters). One wavelength at 432 MHz is approximately 2.28 feet (0.694 meters) and 0.76 feet (0.231 meters) at 1296 MHz. Hence its beamwidth would be approximately 8

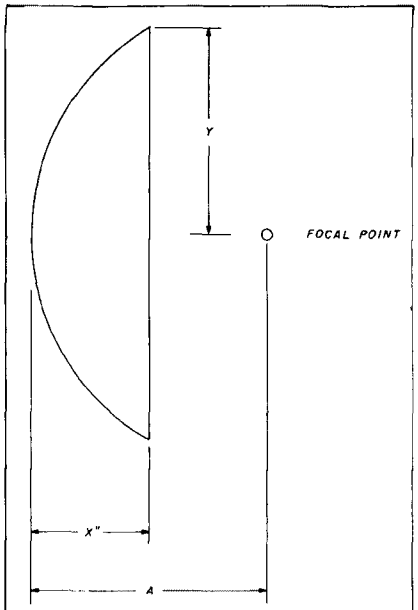


fig. 4. This figure shows the mechanical properties of a parabolic dish reflector.

degrees at 432 MHz and 2.66 degrees at 1296 MHz.

Properly designed, the parabolic dish has a symmetrical beamwidth the same in both E and H planes. The graph in **fig. 3** has been prepared to simplify beamwidth estimation.

design parameters

So how does one go about designing a parabolic dish? What tradeoffs are required? Where do you feed the dish?

First you have to understand the geometry of a parabolic dish. You must have a reflector that follows a parabolic curve. This is described mathematically by **Eqn 5** below:

$$Y^2 = 4AX \qquad (5)$$

where, referring to **fig. 4**, Y is the radius of the reflector, A is the focal length and X is the depth of the reflector, all in the same unit of measure.

If the diameter and focal point (as described below) are known, we can rearrange **Eqn 5** to a more convenient form as shown below:

$$X = Y^2/4A \qquad (6)$$

For instance, suppose that you want to design a 20-foot (6.1 meter) diameter parabolic dish with a focal length of 10 feet (3.05 meters). Using **Eqn 6**, the edge of the dish will be 2.5 feet (65 cm) higher than the center of the dish. Likewise, a point halfway from the center will be 0.625 feet (19 cm) above the center, and so forth. **Table 1** shows a typical parabolic curve for this example. With all the personal computers available, there are many kinds of programs to simplify the computation of a parabolic curve. References 12 and 13 will help you select such a program.

Finally, if the diameter and depth of the reflector are known, the focal point can be found by rearranging **Eqn 5** as follows:

$$A = Y^2/4X \qquad (7)$$

If you already have a dish, the focal point can easily be determined using **Eqn 7**. First measure the reflector diameter with a tape measure. The radius, or Y, is one-half the diameter. If a board or string is placed on the rim from one side of the reflector to the other, the depth, X, can be easily measured. For example, if we have a dish with a depth of 2.5 feet (0.762 meters) and a radius of 10 feet (3.048 meters), the focal point will be 10 feet (3.048 meters).

Next we must determine the f/d or focal length to diameter ratio using the following equation:

$$f/d = A/2Y \qquad (8)$$

Using **Eqn 8**, the f/d ratio in the example above is 0.5 (10/20 feet or 3.05/6.1 meters). We now have all the mathematical parameters necessary to talk intelligently about the design of a parabolic dish type of antenna.

parabolic dish characteristics

So why are the diameter, depth, focal point, and f/d ratio so important? It should be intuitive that the diameter of the reflector is the principal factor that determines the gain of a parabolic antenna. The larger the diameter, the higher the potential gain, as shown on the graph on **fig. 2**.

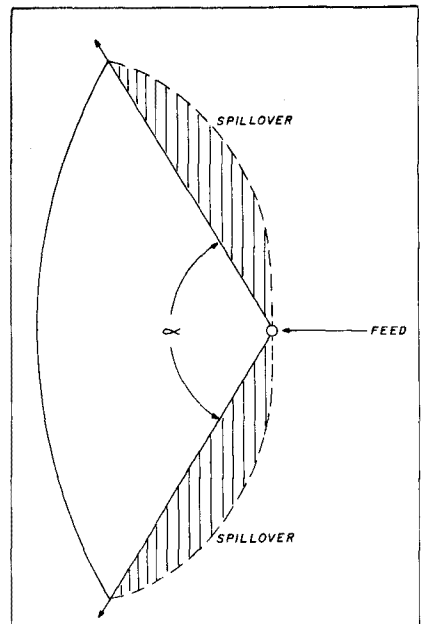


fig. 5. This figure shows how the feed system illuminates the reflector of a parabolic dish. Note that the spillover is the radiation that extends beyond the edge of the dish.

The reflector must follow a parabolic profile (see **eqn. 5**). This curve is calculated after the diameter and f/d ratio of the dish are selected. As dish diameter increases, greater care must be taken to maintain the shape of the reflector.

The depth of the reflector is the parameter that determines the focal length: the deeper the dish, the shorter its focal length. In a "deep" dish, where the f/d ratio is approximately 0.25, the focal point will be on a line parallel to the rim of the dish. The "flatter" the dish, the longer the focal length — and the more difficult it becomes to hold the feed system in place mechanically.

The f/d ratio, however, is probably the most important parameter when designing a parabolic dish antenna from scratch. Let's see why.

In order to use the parabolic dish reflector properly, the feed system must be placed at or near the focal point. Because this feed system "looks" into the dish from the focal point, it must illuminate the subtended

angle between the rims of the dish with a properly tapered radiation pattern as shown in **fig. 5.** (More on this shortly.)

If you have a deep dish with an f/d ratio between 0.25 and 0.4, this subtended angle will be quite large — from 180 to 128 degrees, respectively. If a dish is relatively flat with an f/d ratio between 0.5 and 0.8, the subtended angle will be only 106 to 70 degrees, respectively. For convenience, various subtended angles versus f/d ratios have been plotted in **fig. 6.**

Why is this angle so important? The reasons are many, and all are related to the f/d ratio. Let's first look at some of the tradeoffs.

There are advantages to a low f/d ratio (0.25 to 0.4). This type of dish has a short focal length and hence the support of the feed system is easier from a mechanical standpoint. Generally speaking, the deep dish is quieter on reception. A 0.35 to 0.4 f/d ratio is very common in commercial dishes.

There are also disadvantages to a low f/d ratio. The feed system has to see a wide angle as previously discussed; this makes the feed system more difficult to design. The proximity of the feed to the reflector can cause the feed impedance to change when it's mounted at the focal point. From a mechanical point of view, the deeper the curve, the more difficult is is to fabricate the reflector. There's also greater likelihood that the polarity will become distorted. Usually the low f/d ratio dish demonstrates only moderate efficiency.

A high f/d ratio (0.5 to 0.6) offers several advantages. Because its illumination angle is narrower, the feed system is usually easier to design. The reflector has a flatter profile, making it easier to fabricate. Since the feed point is further from the reflector, it's less likely to be detuned by the reflector. Overall reflector efficiency is generally higher than in a dish with a low f/d ratio.

There are some disadvantages to a high f/d ratio. The feed system, for example, is much further from the reflector and hence more difficult to support. Even if the feed system is properly designed, there will be more spillover and hence more noise pickup than would be expected in a low f/d ratio dish.

When designing a dish, therefore, the main considerations are the mechanical design and the feed system, which are both governed by the f/d ratio as discussed. From our point of view, as Amateurs, an f/d ratio of 0.45-0.6 seems to be the best compromise since the feed system and the surface of the reflector are easier to design.

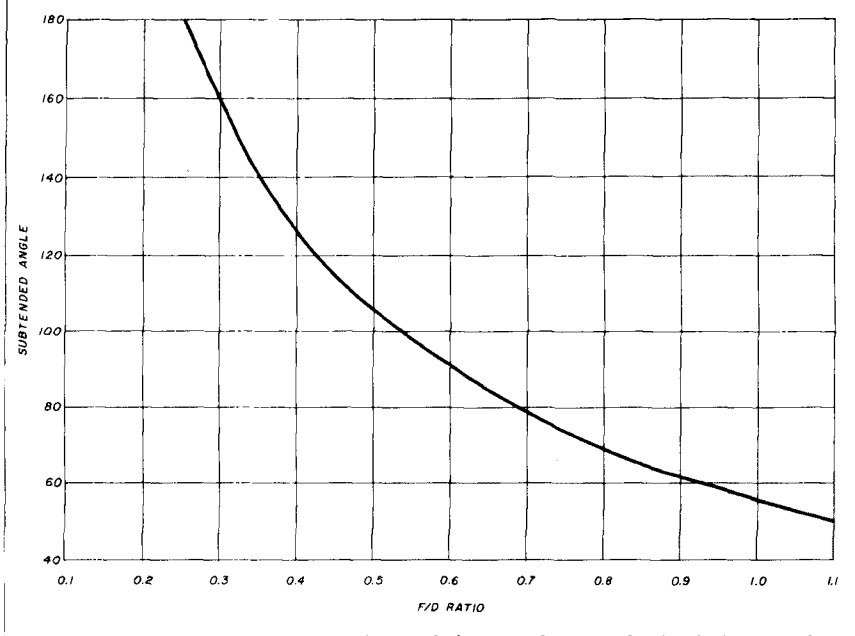


fig. 6. This graph shows the subtended angle (as seen from the feed point) versus the F/D ratio of a parabolic dish reflector.

table 2. Gain and sidelobe levels versus illumination taper on a typical well-designed parabolic dish.

edge illumination in dB	typical level of first sidelobe	approximate overall aperture efficiency (%)
-6	-20.0	68.0
-10	-23.5	81.0
-15	-31.0	77.5
-20	-46.0	69.0

feed systems

So far the feed system itself has been mentioned only in passing. Let's look more deeply into the way it has to perform.

As I said earlier, the feed system must illuminate the subtended angle. Furthermore, a parabolic dish is symmetrical in the horizontal as well as the vertical plane, so the feed system must also be symmetrical in both planes.

Although actual feed systems will be discussed in next month's column, their requirements will be spelled out now. It would appear that the feed system should have a half-power beamwidth equal to the illumination angle as shown in **fig. 6.** However, if this were true, a considerable amount of the feed power would be lost

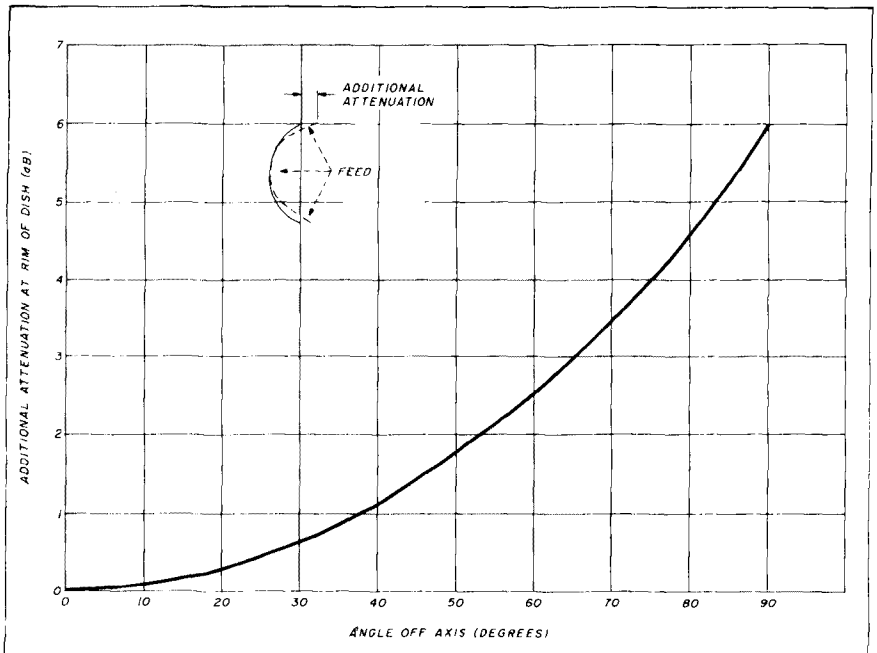


fig. 7. This graph shows the additional "space attenuation" that occurs versus the angle off axis from the feed point on a parabolic dish reflector.

because of "spillover" (see **fig. 5**). Therefore the sidelobes would be high, and in the case of EME, considerable noise would be picked up off the rear of the dish, which is usually the warm earth at 290 degrees Kelvin. On the other hand, if the illuminator feed angle were too narrow, only the center of the reflector would be active and the efficiency of the antenna would decrease — but so would the sidelobes, etc.

Silver[7] and others has discussed this relationship in detail. Most engineers agree that for best overall performance, the optimum feed system should have its −10 dB points right on the rim of the dish. This will yield a reasonably clean pattern and fair gain. The first side lobes will be approximately 23 dB below the main beam. Some typical illumination tapers are shown in **table 2**.

Over-illumination with less than 10 dB feed system taper at the rim of the dish will cause spillover, lower gain, and sidelobes that will be less than 23 dB below the main beam. Conversely, if the dish is under-illuminated and the taper at the rim of the dish is too great, the gain will decrease, as will the side lobes and the noise pickup.

Another factor must be considered: the distance from the reflector to the feed point changes across the reflector. So *you can't assume that* the feed must be optimum just because the feed system has a −10 dB beamwidth equal to the subtended angle. **Figure 7** shows this amplitude taper as a function of the angle as seen from the feed system. Hence this additional factor has to be taken into consideration when choosing the feed system and the f/d ratio. The lower the f/d ratio is, the greater the "space loss" taper at the edge of the dish will be.

summary

This concludes part 1 of our discussion on the subject of reflector antennas. This month's column has described the relationship between electrical and mechanical considerations in designing a reflector-type antennna and, more specifically, a parabolic dish antenna.

It has also been shown that design tradeoffs are primarily linked to the f/d ratio chosen. Next month's column will deal more with the mechanical design of a reflector and the electrical design of its feed system. Once you have this material in hand, you should be able to design your own parabolic dish antenna and know that it *will* work as planned!

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13. Don Cook, K1DPP (see callbook) has a free program listing for Commodore 64 computers. Enclose SASE with request.

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improving the audio on the ICOM IC-27

Make an otherwise fine radio even better

The ICOM IC-27 FM transceiver remains, in my opinion, one of the most compact and versatile mobile units available. The performance of my IC-27A was very impressive, with just one noticeable exception: the received audio was poor. Low volume and poor treble response made many noise-free signals all but unintelligible. Many other Amateurs have experienced this problem as well. Fortunately, there's a very simple remedy.

My first inclination was to experiment with an external speaker. The speaker included in the IC-27 is very small. Often such small speakers are inefficient and have poor response. Yet experimentation with a much larger speaker with a more powerful magnet showed absolutely no discernible difference. The only way one could hope to increase the volume would be by mounting an external speaker closer to the user's ear. This would alleviate only part of the problem, however, since the tonal response would remain virtually unchanged. The latter fact was an important clue, because it led me to believe that the audio amplifier inside the IC-27 was the culprit.

strategy

The solution to this problem is simple. Just change one resistor and one capacitor in the audio section. There's absolutely no reason to fear the operation; it's very simple. ICOM has provided excellent schematics with the radio, the board layout is not terribly compact in this area, and all cables running to the important circuit board are provided with well labeled pull-off connectors in an unambiguous configuration. Aside from forgetting to replace a connector, it's virtually impossible to connect anything incorrectly.

As shown in the ICOM schematic, not reproduced here, Q5 and Q6 form an active filter which seems to effectively filter the intelligibility of many Amateurs' speech. Capacitor C67 was found to form the bottleneck; a smaller value is required here. This greatly improves the treble response.

The audio output power was increased by changing the feedback resistor, R31, of the audio power amp, IC6 (on the ICOM schematic). It was discovered that a larger value would give greater gain while still not introducing objectionable distortion.

modification procedure

Begin the modification by turning the transceiver over (speaker side up), and removing the four screws holding the bottom cover. Carefully place the bottom cover to the side (as shown in **fig. 1**) so as not to strain the speaker wires. The connector for the speaker wires can then be pulled at the circuit board.

The objective is to remove the circuit board in order to gain access to the parts which must be changed. Carefully remove connectors J2, J6, J10, and J12. With your fingers, remove the cover plate of the synthesizer section (it's the large shiny box on the right-hand side). Remove the four screws in the corners of the large circuit board. Remove J1 (the entire circuit board at the rear of the unit). Remove the two heat sink screws on IC5 and IC6. Remove the hex spacer formerly under the rear circuit board.

At this point, the entire circuit board can be lifted out. Carefully pull it out and locate the audio section toward the front on the left side, as viewed from the component side. Locate C67. This is a 0.01 μF capacitor near IC6. Using a solder wick or a solder sucker, remove C67 and replace it with a tiny 0.001 μF capacitor. A small ceramic part will do nicely. Then locate R31. This is a 150-ohm, 1/8 watt resistor near J6. Replace this with a 270-ohm resistor. If possible, use a 1/8 watt part so that it will fit easily in the circuit board. Clean your solder connections and carefully check for solder bridges.

This completes the modifications. Replace the circuit board and reverse the disassembly procedure, being careful to replace all removed connectors.

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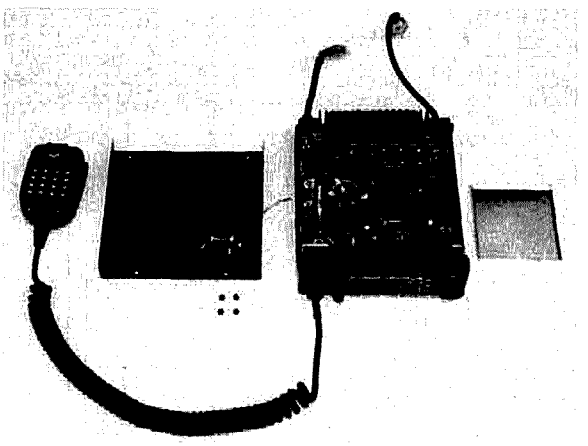


fig. 1. The ICOM IC-27 is easily disassembled for modification.

results

After performing these modifications, I was much happier with the IC-27. Audio volume and clarity are much better than with the original unit, even with the original speaker. In fact, I seriously doubt the need for an external speaker. The internal speaker provides plenty of power, even in my very noisy automobile. I'm sure other Amateurs will agree that this modification makes an otherwise superb radio even better.

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what others think of us

When asked about the place of Amateur Radio in today's rapid technological expansion, academician Vladimir Aleksandrovich Kotel'nikov, Vice President of the Academy of Sciences of the USSR and Director of the Institute of Radiotechnology and Electronics of the Academy of Sciences of the USSR, said (in part), "One of the needs of man is the need to create. Amateur Radio opens up broad possibilities in this connection for youths as well as for adults....

"Amateur Radio has been and remains a wonderful school for mass training of personnel for radio electronics. Many young people entered radio electronics via Amateur Radio, judging by our experience....

"The main goal of Amateur creativity has remained unchanged — to attract youth to active participation in the struggle for technical progress and to teach inventiveness, innovation, and improvements in technology. This is very important for our country."[1] (And for ours, too — Ed.)

quad falling apart?

One of the problems with the multi-element quad antenna is keeping it in the air during the months of bad weather. The point at which the loop wire and support structure meet is an area of particular weakness in this design. If fiberglass or bamboo poles are used, the copper antenna wire will "saw" back and forth at this point under the movement of the element in the wind. This sawing motion often results in the breaking of either the support pole or the wire, and the whole antenna must be taken down for repairs.

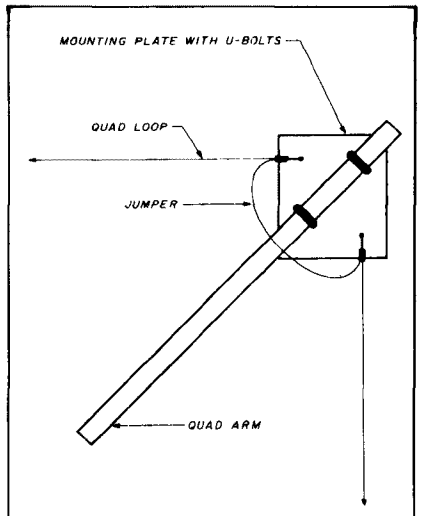


fig. 1. Alternative method of attaching quad loop to quad arm.

Figure 1 shows a novel approach to this problem. An insulating plate (made of fiberglass or formica, for example) is affixed to the quad arm with two U-bolts. The quad wire loop is then attached to holes drilled in the plate and the connection is made as shown in **fig. 1**. Since I don't have a quad up right now, I can't vouch for the idea, but it looks like a good one, and Amateurs planning to erect quads this spring may be interested in trying out this arrangement.

the 160-meter antenna at DJ8WL

DJ8WL, who just returned to Germany after three months in Salt Lake City sent me a sketch of the 160-meter antenna he's using in Germany. Those who've heard his signal can attest to the excellence of the antenna design (**fig. 2**), which he's used since fall, 1984.

His antenna is about 103 feet high (31.4 meters) and is center-loaded by a high-Q coil (L1). A second loading coil is placed at the 19.7 foot (6 meters) level (L2).

Peter uses 50 radials, each 3/8-wave length (205 feet/62.42 meters) long. The feedpoint resistance was measured to be 38 ohms. When only 16 radials were used, the feedpoint resistance was 40 ohms. He estimates his ground resistance to be 9 ohms, so the antenna is about 63 percent efficient.

Antenna response is quite broad, showing less than 1.5:1 SWR over the range of 1800 to 1900 kHz. Peter reports working 9J2JN and LU9EIE the first morning the antenna was on the air!

Peter's antenna was mounted in a tall tree. The top section was made of a fiberglass rod with wire attached, and the lower sections are made of wire. Compared to an inverted-V slung from the same tree at the 80-foot (24.38 meter) elevation, the ground plane produced DX signal reports that were up between 1 to 2 S-units.

DJ8WL's 160-meter DX record with this antenna — as of spring, 1985 — was 135 countries.

aircraft enhancement of VHF/UHF signals

Why does the 2-meter path between San Francisco and Los Angeles pop open every so often, usually in the evening, for just a few minutes? This interesting opening seems to be duplicated in other parts of the world, apparently on a hit-or-miss basis. Doug McArthur (VK3UM) and Ralph Kettle (VK1RK) noticed strange openings on their path, as well as similar openings to VK2ZAB. Path loss was

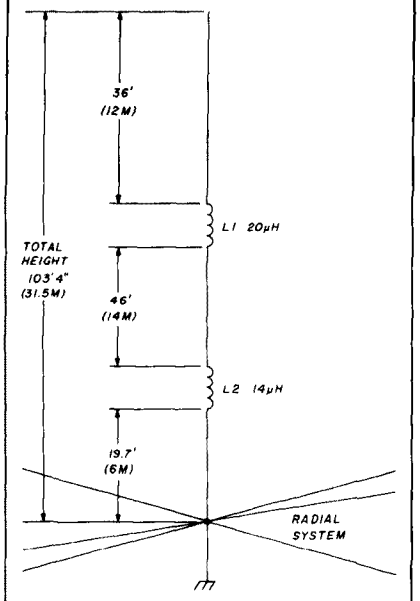


fig. 2. Vertical antenna at DJ8WL has loading coil split in two (L1 + L2). Feedpoint resistance is about 40 ohms.

calculated to be −245 dB. The use of medium power, 36-element arrays, and low noise receivers showed that the paths were open for more time than the calculations would indicate. The Melbourne-Canberra path, in particular, had mountainous terrain along the route so direct (i.e., line-of-sight) communication was out of the question, except for occasional tropospheric enhancement. What was causing the strange VHF propagation?

The time of the opening seemed to be related to aircraft passing between the two stations. In particular, two commercial airliners flying from Sydney to Melbourne were midpath at the times of the enhanced propagation.

Upon investigation, it seemed that the height of the flight path and distance of the aircraft from the stations did not tally with the signals received. The signals were too strong and varied markedly with the weather.

VK3UM and his colleagues in the investigation believe that the aircraft are the cause, but not the actual media of the enhanced propagation.

Possible explanations were reflections caused by the condensation trails ("contrails") left by the aircraft flying at about 30,000 feet, or perhaps refraction caused by the air turbulence wake (temperature effect or vortex turbulence).

The size of the aircraft apparently influenced the enhancement, with the highest level obtained from 747s and DC10s. But even aircraft as small as the Fokker F28 provided a good degee of path enhancement. Multiple aircraft, on the other hand, resulted in multipath effects, making copy difficult.[2]

some good news?

It's reported in the technical literature that the new AN/TPS-118 over-the-horizon radar under construction in Maine will soon be on the air. It will provide radar detection of planes at ranges from 500 to 1800 miles (804 to 2897 km). The device uses a transmitting antenna more than 3600 feet long and radiates up to 100 Megawatts of power (effective radiated power) in the 5 to 28 MHz frequency range. Two other sites are planned, one covering the Pacific Coast and one in the Midwest to fill in the skip zones of the other two. That adds up to *three* of these giants grinding away in the continental United States. Let's hope they're programmed to stay out of the ham bands!

a zig-zag dipole antenna

Most compact dipole antennas make use of some *form of* lumped loading such as an inductor placed in series with the antenna. A few years ago the "zig-zag dipole" was investigated at Hosei University, Tokyo, Japan (fig. 3).

One of the more interesting zig-zag dipoles developed and tested used a 0.5 wavelength wire folded back so that the overall length was about 24 percent less than a full-size dipole. The feedpoint resistance was 46 ohms. The included angle was about 130 degrees and each leg of the configuration was as indicated in fig. 3.

No mention was made of bandwidth, but it seems to me that the antenna configuration may provide improved bandwidth because it resembles a "fat" dipole. Some enterprising experimenter should try this antenna on the 80-meter band and see how it performs.

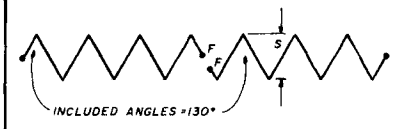


fig. 3. Experimental zig-zag dipole antenna reduces length without use of *loading* coils.

an inexpensive 160-meter antenna that works

No room for a quarter-wave vertical with 120 radials for 160-meter DX work? Too bad. Join me in crying the blues! About the best I can do is a random-length wire running from the house to my nearby antenna tower. I work it against the residence plumbing system as a ground. Recently I added two 8-foot (2.4 meters) ground rods, one at each end of the house, and tied them to the copper water pipe system. So far, so good.

After I'd been working with various end-fed antenna systems for over six months, the back yard was littered with bits of copper wire. I finally ended up with a wire antenna that worked well, although not quite as well as the antennas the "big guns" were using. But if you share my problem of limited space *and* don't want to grow an antenna farm, this antenna is ideal for you.

Basically, it's a wire about 155 feet (47.24 meters) long. This makes it longer than a quarter-wavelength at 1.8 MHz (130 feet/39.62 meters). The feedpoint impedance of the longer wire is inductive and slightly higher than that of the quarter-wave wire. The inductive reactance is tuned out by a series-connected capacitor and the feedpoint resistance of the antenna (about 23 ohms) is transformed by a shunt inductor to 50 ohms (fig. 4).

This design provides an inexpensive matching system to provide a 50 ohm

feedpoint, required by today's modern transceivers. The solid-state amplifiers are not very forgiving with respect to oddball antenna feedpoint impedances, and operation of the 160-meter band stretches the capability of the solid-state amplifier, even under the best conditions.

But this antenna solves that problem. You can change the antenna length and compensate for it with the variable capacitor. As the antenna is shortened toward the quarter wavelength point, the capacitor value increases. As the antenna approaches three-quarter wave resonance, the value of the capacitor diminishes. A happy medium is reached when the antenna is about 0.275 wavelength long. If the inductor and capacitor are variable, the operator can scoot from one end of the band to the other in just a few seconds.

Operational bandwidth of the antenna is quite good; an SWR of better than 1.5:1 can be maintained over 50 kHz without touching the network controls. I tune up at 1825 kHz and am able to QSY up and down 25 kHz at the touch of the transceiver dial.

How well does it work? Well, I'll let you know the results after the next 160-meter DX contest! So far, it seems to be the best wire I've ever had on the "low band."

building the tuner

You can build the tuner on a metal chassis and panel, but it's not necessary (see photo A). My tuner is built on a plywood base with a micarta™ panel. The unit is 8 x 8 inches (25 x 25 x 25 cm). I recommend using a vernier drive on the capacitor because the setting for lowest SWR is quite sharp. The setting of the shunt inductor, on the other hand, is less critical and the counter dial is a luxury. The 3 ampere RF thermocouple meter is very handy for tuneup, so if you can find one at a flea market, buy it and use it. I short mine out with a clip lead after tuneup to keep the meter from jumping around when I'm transmitting.

the 80-40 meter antenna at CN2AQ

Have you heard the strong signal of CN2AQ on 80 meters? He has an antenna that doubles as a telephone line! Look at **fig. 5.** The 328 foot (99.91 meter) wire runs between two buildings. RF chokes at each end of the

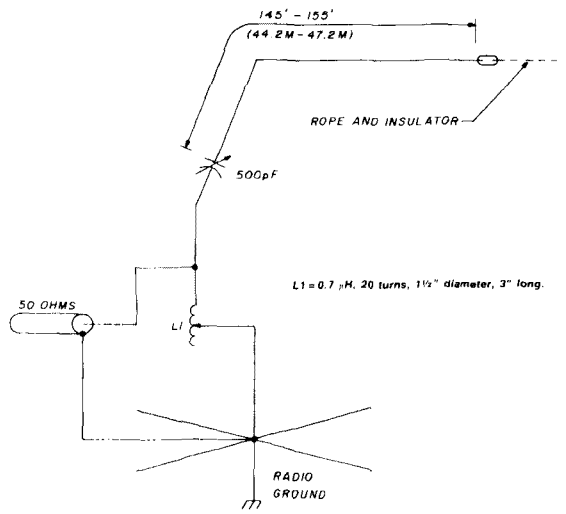


fig. 4. Simple end-fed 160 meter antenna and matching system.

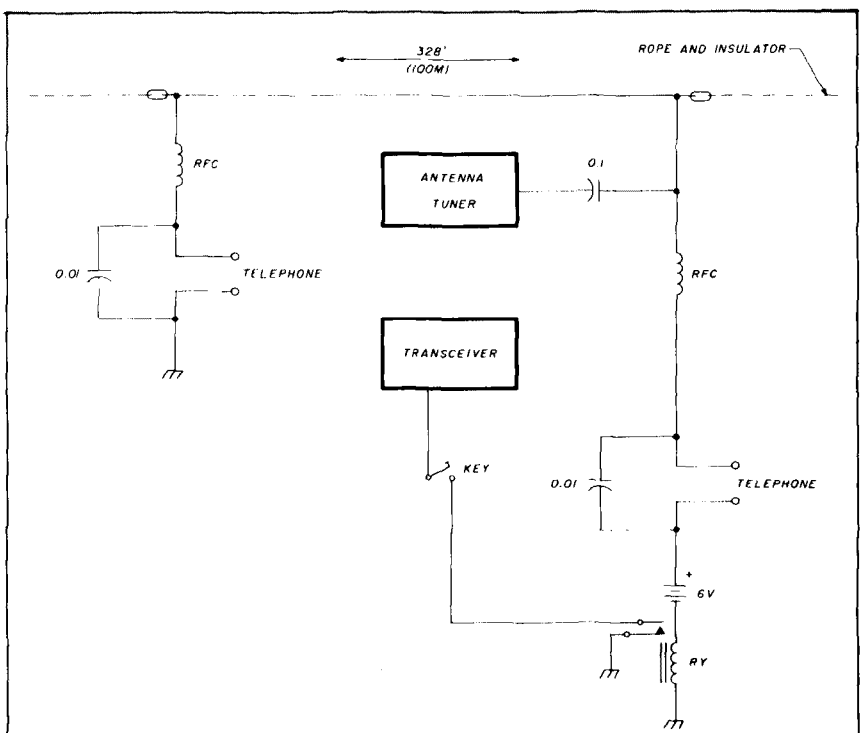


fig. 5. Telephone line and 80-40 meter antenna at CN2AQ. Relay RY opens control circuit of transmitter so it cannot be activated when phone is in use.

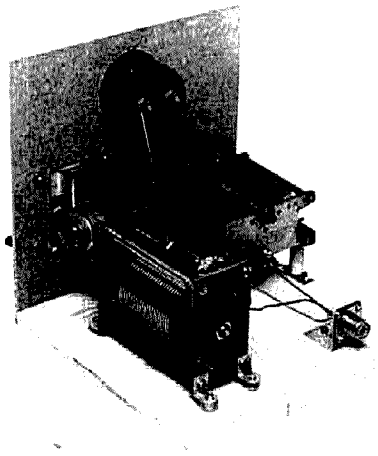


Photo A. W6SAI tuner is built on plywood base.

wire isolate it for the telephone service, in which the return circuit is via ground. In order to prevent the transmitter from being activated when the telephone circuit is in use, a series-connected relay opens the control circuit of the exciter. Now, that seems to be a very practical idea!

what about your antenna?

The sunspot cycle is at low ebb and interest is growing in the 160 and 80-meter bands. I would be interested in hearing about *your* antenna for the low bands. Please send the information to me, care of *ham radio* magazine; the most interesting antennas will be discussed in this column. I hope to hear from you!

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EME link calculator program

Maximize your station's performance with knowledge of tradeoffs

In trying to set up an EME (moonbounce) station I found it difficult to make trade-offs in terms of antenna gain, coax line losses, preamplifier placement, and output power. What I needed was an easy way to weigh the different parameters against each other. Looking through a few books, I found what I was looking for in the *ARRL Handbook*, the *RSGB VHF-UHF Manual* and in an article by R. Lentz.[1,2,3]

By taking pieces of information from all three sources, I was able to figure out how to vary all the station information in order to calculate the link budget signal-to-noise ratio. However, it became apparent that getting answers by punching buttons on my hand calculator would take a few nights. Also, I was mixing up the answers and station data so much that I couldn't figure out what answer belonged to what station data. Figuring that there had to be a better way, I looked for a computer program to do the job. Not finding one, I wrote the BASIC program included in this article (**fig. 1**). I used Microsoft BASIC for the IBM PC, so it should be able to run as is or with little modification on other PCs or compatibles.

After the requested data is entered, the program provides signal-to-noise ratio (SNR), effective isotropic radiated power (EIRP), and the system noise temperature. Using these figures, you can change the input parameters and see how to best assemble your EME system. With the computer doing most of the work, it's easy to use real numbers, rather than using whatever is convenient, as I did with the calculator.

Both the ARRL and the RSGB handbook examples placed the receive preamplifier at the antenna. They then set the line loss at a convenient zero dB. Since no-loss coax phasing lines don't exist, those example calculations don't really show you what is happening in an antenna array. On the other hand, the program shows that even a 0.5 dB phasing line loss will have a significant effect on the performance at 432 MHz. The computer program provides performance information about an EME station while allowing you to easily change all the station parameters.

realistic data is needed

The data requested by the program is generally familiar. The only data I had trouble figuring out was the sky temperature, which is the actual sky temperature the antenna sees. Don't confuse it with the array temperature, which is sometimes given in antenna articles. Try about 10 to 35 degrees Kelvin for 432 MHz and around 175 to 200 degrees Kelvin for 144 MHz.[4] (Use the lower numbers if your system is in the country and the higher numbers if you live in or near an urban area). At low elevations the antenna will see the earth at 290 degrees Kelvin. This program assumes you're looking high enough with your antenna so that the antenna's main and side lobes clear the earth. After a little practice with the program, you can set the sky temperature higher; this will demonstrate earth effects on EME communications at moon rise or moon set. I left the bandwidth as a variable. Most of the time it should be set at 100 Hz, as the ear and mind can provide that much filter capability. You can set the bandwidth to 1800 Hz to see what kind of signal is required for EME SSB.

The program doesn't provide absolute answers. You can't take an existing station, enter its data into the program, and come out exactly with the way the system works. The real world, after all, includes other factors such as Faraday rotation, noise from you don't know where, and systems built not quite the way they should be. The real benefit of the program is to show the relative merit (or lack of merit) of different ways of putting together an EME station. For instance, what's the benefit of running 1.5 kW instead of 800 Watts? How much better is a 0.3 dB noise figure preamp than a 0.6 dB preamp? Does it make any difference if you use Belden 9913 for phasing lines instead of RG-213/U? Try it and see — you may be surprised.

program structure

The program is divided into six parts: constants, data input, calculations, screen output results, print results and program end. **Lines 10 to 35** contain the program header and set up constant values. **Lines 30 and 35** provide a log to base 10 function as most BASICs have only a log to the base e function. **Lines**

By David Engle, KE6ZE, 1063 Summerwood Court, San Jose, California 95132

```
10    ' Program to calculate signal to noise ratios for EME operations.
11    '       David Engle, KE6ZE       September 1985
12    '
20    K = 1.38E-23 ' Boltzman's constant.
25    TL=290.0 ' Temperature of receive feed line.
30    LOG10E=.434294482 'is log to the base 10 of e.
35    DEF FNLOG10(V)=LOG(V) * LOG10E ' Log to base 10 function
100   ' Get data to set up calculations.
110   PRINT" EME Signal to Noise calculator program" : PRINT
120   INPUT" ENTER FREQUENCY, MHz: ",FREQ
130   INPUT" ENTER TX ANTENNA GAIN, dBi: ",GT
140   INPUT" ENTER TX OUTPUT POWER, WATTS: ",P
150   INPUT" ENTER TX LINE LOSS, dB: ",LT
160   INPUT" ENTER RX ANTENNA GAIN, dBi: ",GR
170   INPUT" ENTER RX LINE LOSS, dB: ",LL
180   INPUT" ENTER RX NOISE FIGURE, dB: ",RN
190   INPUT" ENTER RX BANDWIDTH, Hz: ",B
200   INPUT" ENTER SKY TEMPERATURE, Deg K: ",SKY
210   PRINT
300   'Start calculations.
310   'Transmitter power.
320   PO = 10.0 * FNLOG10(P)
340   'Receiver noise level.
350   TR=((10.0^(RN/10.0)) -1.0 ) * 290.0 'Noise figure to noise temp
360   LR = 10.0^(LL/10.0) ' Noise figure (dB) to noise factor.
370   TS = SKY + ((LR - 1)* TL) + (LR * TR)
380   PN = 10.0 * FNLOG10(K*TS*B)
390   'Path loss.
400   PL = 8.72 * LOG (432/FREQ) - 261
410   'Signal to noise.
420   SNR = PO - LT + GT + PL + GR - PN
430   BAND% = INT (FREQ)
440   EIRP =INT (P*10^((GT-LT)/10.0))
500   'Display the results.
510   PRINT"Transmit EIRP =";EIRP;"Watts,";
520   PRINT" System Noise Temp =";(INT (TS));"Degrees K."
530   PRINT" Signal-to-noise ratio at:"," Perigee","Mid.","Apogee."
540   PRINT" for ";BAND%;"MHz band is:", SNR, SNR-1, SNR-2
550   PRINT
600   'Print the results.
610   OPEN "LPT1:" FOR OUTPUT AS #1
620   PRINT#1,
630   PRINT#1,"FREQUENCY =";FREQ;"MHz"
640   PRINT#1,"TX ANTENNA GAIN =";GT;"dBi,";
650   PRINT#1," TX OUTPUT POWER =";P;"Watts,";
660   PRINT#1," TX LINE LOSS =";LT;"dB"
670   PRINT#1,"RX ANTENNA GAIN =";GR;"dBi,";
680   PRINT#1," RX LINE LOSS =";LL;"dB"
690   PRINT#1,"RX NOISE FIGURE =";RN;"dB,";
700   PRINT#1," RX BANDWIDTH =";B;"Hz"
710   PRINT#1,"SKY TEMPERATURE =";SKY;"Deg K"
720   PRINT#1,
730   PRINT#1,"Transmit EIRP =";EIRP;"Watts,";
740   PRINT#1," System Noise Temp =";(INT (TS));"Degrees K."
750   PRINT#1,"Signal-to-noise ratio at:"," Perigee","Mid.","Apogee."
760   PRINT#1," for ";BAND%;"MHz band is:", SNR, SNR-1, SNR-2
770   PRINT#1,
780   CLOSE #1
900   INPUT" Do you wish to continue? (Y/N) ",ANS$
910   IF ANS$="Y" OR ANS$="y" THEN 120
```

fig. 1. EME station optimization program listing.

```
Frequency = 144 MHz
TX Antenna Gain = 21 dBi, TX Output Power = 1500 Watts,
                                    TX Line Loss = 0.5 dB

RX Antenna Gain = 21 dBi, RX Line Loss = 0.5 dB

RX Noise Figure = 0.5 dB, RX Bandwidth = 100 Hz

Sky Temperature = 170 Deg K

Transmit EIRP = 168302 Watts, System Noise Temp = 245
                                              Degrees K.

Signal-to-noise ratio at:  Perigee,   Mid.,     Apogee
for 144 MHz band is:       6.548783   5.548783  4.548783
```

fig. 2A. Typical data displayed in run of SNR program.

```
PO = 31.760
TR = 35.385
LR = 1.122
TS = 245.088
PN = -184.708
PL = -251.420
SNR = 6.548
```

fig. 2B. Intermediate values generated for fig. 1 results.

100 to **210** inquire and retrieve the station data from the operator (you). **Lines 300** to **440** do the calculations and come up with the answers. **Line 350**, which converts the preamplifier gain in dB to noise temperature, is from the RSGB manual. **Line 360** converts the receiver (preamplifier) line loss into a ratio.[3] **Line 400** provides the path loss. (The path loss is an approximation — a little high at 50 MHz, right on at 432 MHz, and a little low at 2304 MHz. All are within 1 dB.) **Lines 500 to 550** display the answers on your screen. **Lines 600 to 780** print both the input data and the answers on your line printer. If you have no printer, omit **lines 600** to **780**. **Line 610** sets the printer up. If your computer calls its printer something other than "LPT1," you'll have to change this line accordingly. **Lines 900** to **999** re-run or end the program.

sample run checks program

Figure 2A contains a set of data to run through the program after you get it entered in the computer. This can be used to check the accuracy of what you've typed in. If you don't get the answers listed to within 0.1 dB, exit the program by answering "N" to the question, "Do you wish to continue?" You should still be running BASIC at this point, so you can look at the intermediate answers from the program. Simply type "PRINT PO," "PRINT PL," and those values will be printed on the screen. Any of the intermediate values used in the program (e.g., PO, TR, LR, TS, PN, PL, SNR) may be looked at in this manner. **Figure 2B** contains the set of intermediate values generated during the **fig. 2A** example. If your answers are different from these examples, then you have an error in one of the previous lines that works on that variable.

I hope this program can help you demystify the optimization of your EME station. A couple of the more interesting things I noticed were the effects of phasing line loss and preamplifier noise figure on the system's performance. These effects are easily noticed in the system noise figure. Fire it up and plug in your numbers. It may be easier to improve your system than you think!

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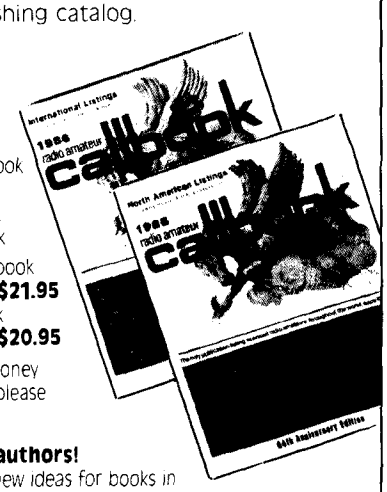
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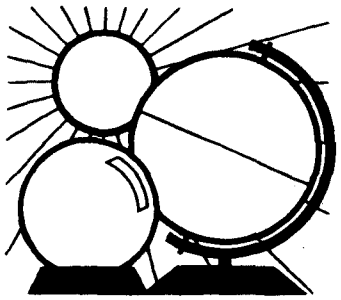
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DX FORECASTER

Garth Stonehocker, KØRYW

sunspots and MUFs

Because we're nearing the 11-year sunspot number minimum, the Institute for Telecommunication Science study comparing MUF distribution of values for low, medium, and high sunspot numbers should be of interest. **Table 1** (similar to last month's) lists *the ratios of the 10- and 90-percentile* monthly median MUF values. The data is separated into day and night groups over three solar activity ranges: low (68-88) sunspots in the years 1954-55; medium (90-115) in the years 1961-62; and high (200-285) in the years 1957-58 at various control point latitudes. The upper decile values (90 percent) use the column head + % and the lower decile (10 percent) by – % in the table. A separate listing is provided for daytime and nighttime periods.

Notice that greater deviations from median occur in the high (70 degree) and low (10 degree) latitudes in all sunspot categories regardless of whether it's local daytime or nighttime. However, the largest deviations occur during a high sunspot number year and at night. This increased MUF is from the solar flare particles arriving via the solar wind at the polar ionosphere down the open geomagnetic field lines from the sun (see the geomagnetic field drawing in "DX Forecaster," June, 1985, page 102) on the day side of the earth. On the night side the particles come into the ionosphere from the stored particle areas (Van Allen belts) along the closed and tail geomagnetic field lines.

The exception to the foregoing pattern is the 20-degree latitude night values at low and medium sunspot numbers. Here the geomagnetic variations help the ions drift upward along the closed field lines until within 20 degrees of the geomagnetic equator. This is a winter nighttime effect during medium to low sunspot numbers. Here again is evidence of trans-equatorial DX propagation; it is accompanied by an F-region trough (with low MUFs) just south of the auroral zone. The low MUFs in the trough are shown by the large – % numbers at 70 degrees latitude during the night. The increased geomagnetic variability during low to medium sunspot numbers results from a stronger solar wind stream through a thinning solar corona while the sun is yet somewhat active. This data, the MUF distribution statistical trends of propagation effects, can be used to understand and improve DX operating skills.

last minute forecast

The best DX conditions for the higher frequency bands are expected during the first and last weeks of February. The solar flux base level should increase about 5 units in January and February. So look for some increase in MUFs in the midlatitudes as last month's table showed. Trans-equatorial propagation should be good and enhanced by some geomagnetic disturbances around the 10th and 24th of the month. The lower frequency bands are expected to be very good

table 1. Variation in the ratio of monthly median MUF values.

Latitude	Sunspot Range	– % Day	+ %	– % Night	+ %
70	Low	22	23	28	29
	Medium	24	29	23	30
	High	26	29	27	30
50	Low	19	15	21	21
	Medium	21	17	21	21
	High	22	22	21	18
30	Low	17	17	22	26
	Medium	19	17	24	25
	High	14	17	18	18
20	Low	21	19	26	31
	Medium	18	19	27	34
	High	12	14	21	26
10	Low	14	18	26	27
	Medium	15	17	25	27
	High	12	15	20	25

GMT	PST	WESTERN USA N	NE	E	SE	S	SW	W	NW	MST	MID USA N	NE	E	SE	S	SW	W	NW	CST	EST	EASTERN USA N	NE	E	SE	S	SW	W	NW
0000	4:00	30*	40	20	10	12	10	10	20*	5:00	40	40	20	10	12	10	10	20	6:00	7:00	40	40	20	12	12	12	12	30
0100	5:00	20	40	20	10	12	10	10	20	6:00	30	40	20	12	12	10	12	20	7:00	8:00	40*	40	20	15	12	15	15	30
0200	6:00	20	40	20	12	12	10	10	20	7:00	30	40	20	15	12	12	15	30	8:00	9:00	40	40	20	15	15	20	20	40
0300	7:00	30	40	20	15	12	10	12	20	8:00	30	40	20	15	15	15	20	30	9:00	10:00	40	40	20	20	20*	20	20	40
0400	8:00	30	40	20	20*	12	15	15	30	9:00	40	40	20	20	20*	15	20	30	10:00	11:00	40	40	20	20	20	20	20	40
0500	0:00	40	40	30	20	15	15	20	30	10:00	40	40	30	20	20	20	20	40	11:00	12:00	40	40	30	20	20	20	20	40
0600	10:00	40	40	30	20	20*	20	20	40	11:00	40	40	30	20	20	20	20	40	12:00	1:00	40	40	30	20	20	20	20	40
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0900	1:00	40	40	30	20	20	20	20	40	2:00	40	40	30	20	20	20	30	40	3:00	4:00	40	40	30	20	30	30	30	40
1000	2:00	40	40	30	20	20	20	20	40	3:00	40	40	30	20	30	30	30	40	4:00	5:00	40	30	15	30*	30	30	30	40
1100	3:00	40	40	30	30	20	20	20	40	4:00	40	30	30	30	30	30	30	40	5:00	6:00	40	20	12	20	30	30	30	40
1200	4:00	40	40	30	30	30	20	30	40	5:00	40	20	15	30	30	30	30	40	6:00	7:00	40	20	10	15	20	30	30	40
1300	5:00	40	30	15	20	30	30	30	40	6:00	40	20	12	20	20	30	30	40	7:00	8:00	40	20	10	12	15	20	20*	40
1400	6:00	40	20	12	15	30	30	30	40	7:00	40	20	10	15	20*	20	20	40	8:00	9:00	40	20	10	12	15	20	20	40
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1800	10:00	40	20	10	10	12	20*	20	40	11:00	40	20	10	10	12	15	20	40	12:00	1:00	40	20	10	10	12	15	20	40
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FEBRUARY		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN			ASIA FAR EAST	EUROPE	S. AFRICA	CARIBBEAN S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.

*Look at next **higher** band for possible openings.

throughout the month, but especially good around mid-month.

No significant meteor showers occur during February. The full moon is on the 2nd and lunar perigee is on the 4th.

band-by-band summary.

Ten, twelve, fifteen, and twenty meters will be open from morning to early evening almost every day, and to most areas of the world. The openings on the higher of these bands will be shorter — to the southern hemisphere — and will occur closer to local noon. Transequatorial propagation on these bands will more likely occur toward evening during conditions of highest solar flux and a disturbed geomagnetic field.

Thirty and forty meters will be useful almost 24 hours a day. Daytime conditions will resemble those on 20 meters, except that skip distances and signal strength may decrease during midday on days that coincide with the higher solar flux values. Nighttime DX will be good except after days of high MUF conditions and during geomagnetic disturbances. Look for DX from unusual places on east, north, and west paths during daytime and southern paths at night. The usable distance is expected to be somewhat less than 20 in daytime and greater than on 80 at night.

Eighty and one-sixty meters will exhibit short-skip propagation during daylight hours and lengthen for DX at dusk. These bands follow the darkness regions opening to the east just before your sunset, swinging more to the north-south near midnight and ending up in the Pacific areas during the hour or so before dawn on the path of your interest. The 160-meter band opens later and ends earlier than the 80-meter band.

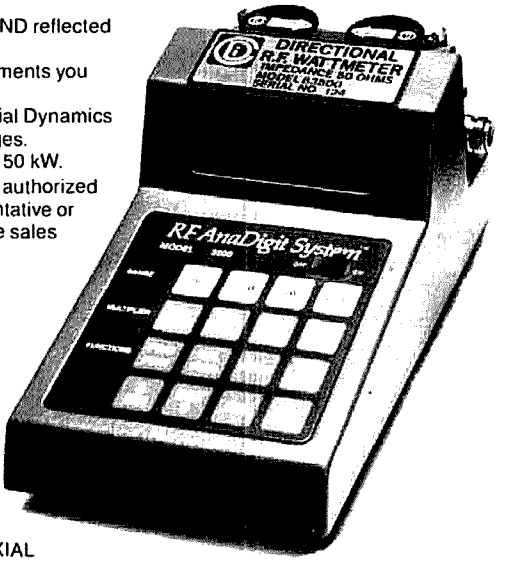
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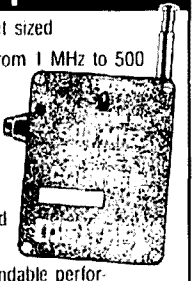
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SAY YOU SAW IT IN HAM RADIO

quartz crystal resonators

Learn how to specify crystals using frequency/temperature relationships

This article provides tools for specifying, testing, and controlling the parameters of piezo-electric crystals. This data is designed to allow you to specify your requirements exactly to the crystal supplier. Also included are TI-55-II calculator programs showing examples that reduce the burden of calculations.

The article is based on practical experience with thousands of crystals used in the communications industry. Because of the competitive nature of the business and the large quantities involved, crystals must be both low in cost and usable, in that they must be capable of being placed on the specified frequency and remain on frequency (within the limits specified) despite changes in both time and temperature.

the crystal impedance meter

When crystals are ordered from catalogs, problems frequently arise. Invariably the crystals quoted are to a military specification. There is nothing wrong with this as long as the circuit in which the crystal has to work matches it and provides the exact conditions specified. In practice this is generally not the case, and as a result, crystals that arrive may be within the specifications when tested in the standard Crystal Impedance (CI) meter, but may not meet your requirements when used in the actual oscillator circuit.

A quick look at some typical crystals listed in a vendor's catalog can give us some clues. We find among the specifications a variety of capacitances quoted. The values shown are frequently relics from the days of vacuum tubes and are used in the CI meter, selectable by a switch. This is all very convenient; the CI meter, however, looks at the crystal and measures it as an impedance — hence its name, "crystal impedance" meter. It is therefore not operating under the conditions that an oscillator places on it. You will also note that some crystals are series resonant. In this case, there is no loading capacitor, because the crystal is being *driven and is* operating into a purely resistive load. These crystals are thus on frequency only if used under those conditions.

As we go above 20 MHz, we get into the realm of overtone crystals. These crystals are, however, not an exact third or fifth or higher multiple of the fundamental. If the use of multipliers is to be avoided, these crystals can serve a very useful purpose, but there are advantages and disadvantages that must be considered.

example using 10-MHz crystal

We will now look at a typical requirement and consider a crystal that operates at 10 MHz in the parallel or antiresonant mode. (The series configuration is referred to as the resonant mode.) The crystal will therefore operate with a loading capacitance of 20 pF and in the fundamental mode. The crystal cut will typically be AT-cut, operating temperature 78.8 degrees F (26 degrees C) and frequency tolerance ±20 ppm (other specifications we will consider later.) After obtaining samples from three different vendors and measuring them in the CI meter, we found each crystal to be at 10.000150 MHz, thus 15 ppm high. All crystals meet specifications, appear to be identical. However, all are not identical because the compliance, C_S has values of 0.01, 0.02, and 0.04 pF, respectively. (C_S, L, and R *are inherent in the* crystal.) **Figure 1** shows the electrical equivalent; this is the crystal without any electrodes attached. Inductance L is the electrical equivalent of the crystal mass. Capacitor C_S is the compliance and R is due to the losses by mechanical friction of the crystal. This electrical circuit is series resonant. **Figure 2** shows the crystal with its deposited electrical contacts. This is achieved by placing a small spot of a conducting material, such as silver, on each

By Peter Boddaert, 11 Blossom Road, Braintree, Massachusetts 02184

table 1.

C_S (pF)	C_L (pF)	$C_S/2C_L$ (pF)	F_S (MHz)	F_P (MHz)	ΔF (Hz)	L (mH)
			crystal 1			
0.01	31	0.0001613	9.998243	9.999855	-145	
0.01	26	0.0001923	9.998243	10.000166	+166	25.339
0.01	21	0.0002381	9.998243	10.000624	+624	
total pulling = 769 Hz = ±38.5 ppm						
			crystal 2			
0.02	31	0.0003226	9.996310	9.999535	-466	
0.02	26	0.0003845	9.996310	10.000154	154	12.674
0.02	21	0.000476	9.996310	10.001068	1068	
total pulling = 1533 Hz = ±76.665 ppm						
			crystal 3			
0.04	31	0.0006452	9.992487	9.998934	-1068	
0.04	26	0.0007692	9.992487	10.000174	+176	6.342
0.04	21	0.0009524	9.992487	10.002004	+2,004	
total pulling figure = 2,004 + 1066 = 3070 Hz = ±153.5 ppm						

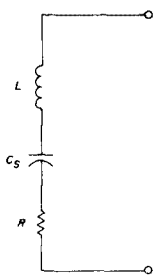


fig. 1. Crystal blank showing relationship of parameters inherent in the crystal: compliance, C_S; electrical equivalent of the crystal mass, L; and losses caused by mechanical friction of the crystal, R.

side and making electrical connections to these. As a result a capacitor is now formed, which together with the holder C, appears in parallel with the series resonant circuit, producing a parallel or antiresonant circuit. This capacitance is shown as C_P with a typical value of 6 pF. C_L is the load capacitance of 20 pF. The crystal thus sees the sum of C_P and C_L = 26 pF as a total load capacitance. In a practical circuit C_L is usually made variable so that the oscillator can be adjusted to frequency. A trimmer capacitor typically provides a minimum variation from 15 to 25 pF. Thus the actual load capacitance can vary between 21 and 31 pF.

We will now calculate the pulling figure or pulling ability of each crystal using the above values of loading C for each of the three crystals. We will also calculate the series resonant frequency, F_S, and the value for L. (See **table 1.**)

A convenient formula for parallel-resonant frequency is

$$F_P = F_S \ [1 + (C_S/2C_L)] \qquad (1)$$

We see that the pulling ability or the ability to place the oscillator on frequency, is greatly influenced by C_S. Failure to control its value can result in being unable to place units on frequency without making modifications or, if the C_S is large such as in crystal 3, the pulling ability can become excessive. Such crystals are prone to frequency jump more so than those that have a smaller value of C_S. It must be remembered that *all* crystals have spurious responses not too far removed from the desired frequency. Such spurs can show up under certain conditions such as when a change in temperature occurs. A simple rule of thumb is: a crystal has a maximum pulling ability of 100 ppm.

We find that crystal 2 meets our requirement of being able to set it to frequency with a sufficient amount of overlap at each end. Crystal 1 would also be suitable; however, at the low end, things are tight. Allowing for component variations in production, together with the fact that the crystal frequency could be higher by another 50 Hz and still be within specifications, it would be desirable to be able to reject those at the incoming inspection. A simple solution to this problem is to specify pulling ability rather than the loading capacitance. The crystal specification would include:

- The frequency.
- Pulling ability in the test circuit (minimum = ±35 ppm, maximum = ±85 ppm).
- Approximate loading capacitance (*for reference only*), 20 pF.

There is, therefore, no reference to frequency tolerance because this is now taken care of by the pulling ability. In addition, the value of the loading capacitance is now used as a general guideline rather than an absolute value.

Of course, the vendor must now be supplied with a test circuit. This will consist of the complete oscillator plus all of its associated circuitry; an isolating stage, mixer, and such. It must be identical with the production units. Keep in mind that any test equipment such as a frequency counter connected to the circuit will have some influence on it, so the coupling must be as loose as possible to minimize these effects. Also be aware that any tuned circuits may have some effect on the frequency. Before shipping the test circuit to your supplier, check for the correlation between it and the in-house standard. There are bound to be some variations between them, and allowance must be made for this.

the overtone crystal

As mentioned, the overtone crystal operates on an odd harmonic of its resonant frequency such as third, fifth, seventh and higher. These crystals are more frequently employed in the series-resonant mode (although they can, of course, also operate in the parallel mode). C_S is extremely small. For example, in a fifth overtone crystal, C_S is on the order of 0.0005 pF at 70 MHz. Hence these crystals are much more difficult to pull on frequency by using a capacitor and consequently, are relatively unaffected by changes in the rest of the circuit because of temperature variations. With these crystals, however, a test circuit should be used and supplied to the vendor.

drive level

In general, crystal levels should be kept low. Some manufacturers specify a maximum level for AT-cut crystals operating in the fundamental mode, as 5 milliwatts. But whatever the maximum level quoted, it is better and safer to stay well below that value for a number of reasons:

- Excessive drive levels can break the crystal.
- The frequency stability is poor because of crystal heating, which causes excessive drifting.
- The crystal is much more susceptible to jump frequency or to operate on one of the crystal spurs, or to deliver an output at two or more frequencies simultaneously. These effects are most likely to show up with temperature variations.
- Note that the drive level will have some effect on the frequency; generally, doubling drive level will cause 2 ppm change.

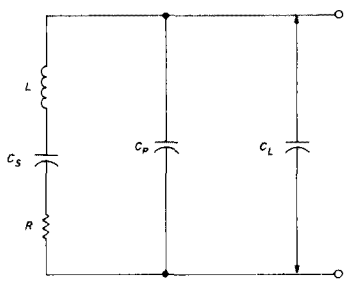


fig. 2. Crystal mounted in holder. C_S; is the loading capacitance.

frequency-temperature characteristics

The AT-cut crystal has been cut at an angle of 35 degrees. **Figure 3** shows the frequency/temperature/angle-of-cut characteristics. Note that the turnover points are placed almost symmetrically on either side of the center point, 79 degrees F (26 degrees C). For example, with the cut at +20 minutes, the upper turnover point (UTP) is at 223 degrees F (106 C) and the lower turnover point (LTP) is at −47 degrees F (−44 C). Thus at 144 degrees F (80 C) and −126 degrees F (−70 C), they are on either side of the center point.

At times, then, one would specify these points — for example, if the crystal were to be placed in an oven that operates at, say 176 degrees F (80 degrees C). It would then be a good policy to have the crystal *on frequency at the turnover point* of 176 degrees F (80 C), for at that point the rate of change is least, thus improving frequency stability. Bear in mind, however, that components that are not temperature controlled, but form a part of the frequency-determining network, will still have an influence on frequency stability; thus, if at all possible, they too should be inside the oven.

Another frequently called specification states that the frequency must be within ± 10 ppm over the temperature range of 158 degrees F (70 C) to −22 degrees F (−30 C). Inspection of **fig. 3** shows that the cut at +4 minutes can meet this, but only barely. There is no margin for error in the angle of cut or for external component changes or tolerances. As a consequence, in the test area, the rejection rate is likely to be high. Generally, oscillators of this nature incorporate some form of temperature correction circuit consisting of thermistors or varactor diodes.

Unfortunately, these circuits invariably have to be adjusted during test, thus causing an added expense.

using NTC capacitors

Figure 4 shows a typical oscillator. Assuming the capacitances due to the transistor are small and can be ignored, we have a 20 pF loading capacitance made

table 2. Results.

temperature	−2.0 F (−19 C)	25 F (−4 C)	52 F (11 C)	79 F (26 C)	106 F (41 C)	133 F (56 C)	160 F (71 C)
F_P (MHz)	9.999804	9.999879	9.999954	10.000031	10.000109	10.000189	10.000270
ΔF (ppm)	−19.550	−12.137	−4.589	3.045	10.914	18.866	26.951
C_L (pF)	21.755	21.059	20.385	19.732	19.101	18.490	17.899

table 3. Corrected figures.

temperature	−2.0 F (19 C)	25 F (−4 C)	52 F (11 C)	79 F (26 C)	106 F (41 C)	133 F (56 C)	160 F (71 C)
ΔF (ppm)	−22.595	−15.182	−7.634	0	7.869	15.821	23.906

up by the trimmer in series with the two 150-pF capacitors. Most mobile radio equipment is required to operate over the temperature range of −22 degrees F (−30 degrees C) to 158 degrees F (70 C). Thus with a total temperature change of 180 degrees F (100 C) the N2200 capacitors will change by: TC × change in degrees F × 5/10,000 × 9 equals percent. (TC × change in degrees C/10,000 = percent), where TC is the temperature coefficient. Thus 2200 × 100/10,000 = 22 percent, so that the total value of the two 150-pF capacitors in series = 75 ± 11 percent. At −22 degrees F (−30 C) the capacitance is 83.25 pF; at 158 degrees F (70 C) it is 66.75 pF.

The trimmer, which has a value of 27 pF, will change by 7.5 percent over the temperature range, thus at −22 degrees F (−30 C) it will be 28.01 pF and at 158 degrees F (70 C) it will be 25.99 pF. The total C that the crystal thus works into at 158 F (70 C) is the series combination of 66.75 + 25.99 pF in parallel with a C_P of 6 pF, or a total of 18.71 + 6 = 24.71 pF. At −22 degrees F (−30 C) this becomes 20.96 + 6 pF = 26.96 pF.

If we use these capacitors in conjunction with crystal 2, which has a C_S of 0.02 pF and F_S = 9.99631 MHz, we thus obtain F_P at 158 F (70 C) = F_S × 1.0004048 = 10.000357 MHz, and at −22 F (−30 C) = F_S × 1.0003709 = 10.000018 MHz. The total frequency change thus brought about = 339 Hz, a variation of 33.9 ppm. With the help of **fig. 3**, we can now locate a crystal that has its turnover points at ±17 ppm. The chart shows an 8 minutes cut to be close, with turnover points at −11.2 degrees F and 168.8 degrees F (−24 C and 76 C, respectively) and a frequency variation of + and −20 ppm. The frequency changes due to the crystal would thus be reduced to approximately ±3 ppm by the use of NTC capacitors, a very worthwhile improvement obtained without any increase in cost or number of components. Of course, the example given here is not the only combination. Had we chosen crystal 3, with a C_S of 0.04 pF, we would have obtained twice as much pulling ability, and thus use capacitors with a smaller NTC. Using a 12 minute angle places the turnover points further out thus extending the usable temperature range of the oscillator. Do not, however, lose sight of the fact that any circuit variations brought about by temperature changes will have an effect on the frequency so that bypass capacitors, layout and component placement, and good engineering principles must be taken into account.

long-term stability

Some manufacturers publish figures for their crystals. In general the cold weld type holder will have better long-term stability. It is not easy to determine as it is a matter of time; however, to achieve the greatest stability the oscillator should operate continuously. It should also be subjected to a program of temperature cycles with the oscillator operating before setting it to frequency. How much aging can one expect? It is not unreasonable to look for a ±3 ppm/year figure, which in most applications is very acceptable.

resistance and temperature

The series resistance of the AT-cut crystal tends to go down in value as the frequency increases when operating in the fundamental mode. With overtone-mode crystals the reverse occurs. In the range of 5 to 20 MHz, a maximum value of 50 ohms is reasonable. Overtone crystals tend to have series R values about 50 percent higher. The resistance tends to be temperature sensitive and can easily double in value at 158 F (70 C). In some cases this could lead to oscillator failure or to a considerable reduction in output power. Specifying the maximum R over the operating temperature range rather than at 79 F (26 C) can avoid any misunderstandings with the manufacturer over this matter. Keep in mind, however, that over-specifying costs money. The tighter the specifications, the

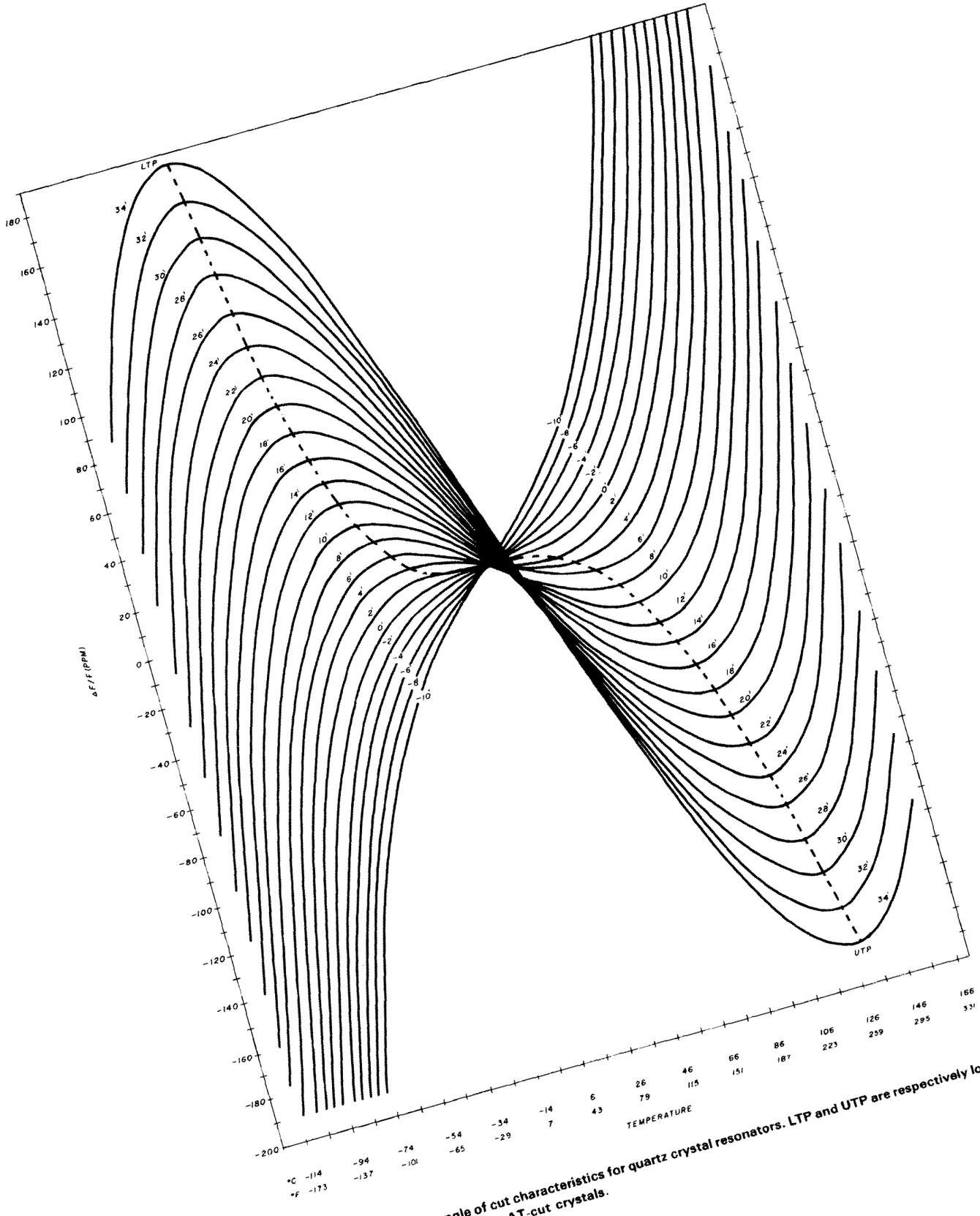


fig. 3. Frequency/temperature angle of cut characteristics for quartz crystal resonators. LTP and UTP are respectively lower and upper turnover points. Chart is for AT-cut crystals.

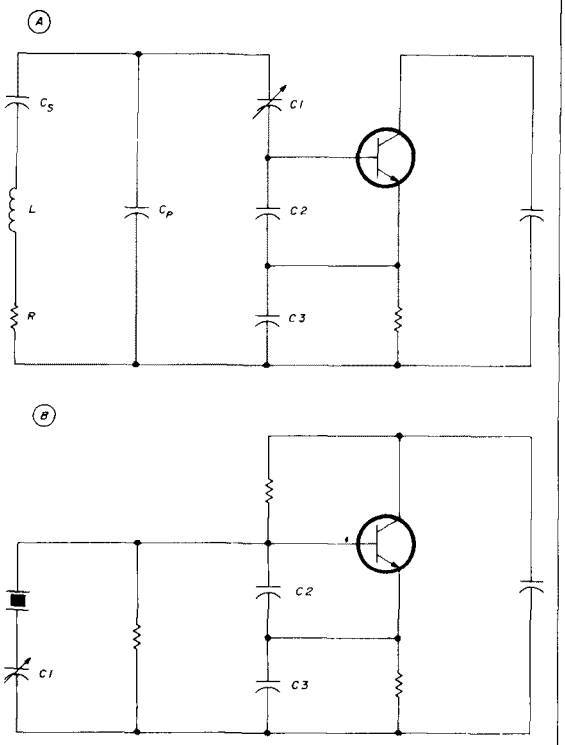


fig. 4. (*A*) is a typical oscillator with a 20 pF loading capacitance made up by the trimmer, C1 in series with two 150 pF capacitors, C2 and C3. In (*B*) a practical circuit in the parallel-resonant mode is shown. Trimmer C1 now has its rotor grounded, making it cold in terms of RF. Bias resistors are also shown.

more expensive the product. Very low series R is also possible, which in some cases can lead to oscillator squegging. Placing a lower limit on the R value can avoid this problem.

spurious responses

As mentioned earlier, all crystals have a number of resonances at unwanted frequencies and in some instances located close to the desired frequency. In some situations this can present a problem, with the oscillator suddenly jumping to one of these. A consultation with the vendor on this matter and a thorough testing of the initial design for this potential trouble is most desirable. If necessary the spurious response characteristics of the crystal should be included in the specifications. Experience shows that these precautions can save a lot of headaches later on.

vendor selection

The vendor is most important; good rapport is essential. Make it clear at the onset that you will not accept units that fail to meet the specifications and that he owns *all* rejects no matter how large the quantity. Also let him know that you intend to conduct inspection tests for compliance to specifications. For small quantities, every unit will be tested; for large quantities testing will be conducted on a sampling basis. The sampling method plus the amount of rejects found in the sample size before rejection should be explained and agreed upon by all concerned, because this will be part of the purchase contract.

Crystal manufacturers tend to specialize; consequently, the vendor who makes excellent low-frequency crystals may not produce the best high-frequency crystals. This can be especially true for overtone crystals so make sure that you select a vendor who demonstrates both the ability and willingness to produce crystals that consistently meet all your requirements. It is a mistake to press a supplier to make something that he is not really comfortable or experienced with; sooner or later, things will go wrong, spoiling a good relationship.

Today, computers and software that will determine the pulling ability of the crystals and print out a histogram of the results as well are easily available. A copy of this information sent to the supplier will enable him to make any slight modifications to his next production run if necessary.

determining F_P, ΔF, C_L

We have looked at quartz crystals in general and at some of the desirable parameters. I also proposed the use of NTC capacitors to compensate the frequency deviation of the crystal with temperature. I then showed that pulling ability is important, much more so than is generally realized. By placing a minimum/maximum limit on it, we get the Series C, a property of the crystal under manufacturing control, ensuring consistency of the product as well as the ability to place the crystal on frequency in our circuit and to apply temperature compensation.

In the previous example the minimum/maximum pulling ability was specified as ± 35 ppm and ± 85 ppm. Statistically the mean is thus ± 60 ppm. Inspection of the formula for F_P and the example crystals 1, 2, and 3 shows that the pulling ability of the crystal is directly proportional to the value of C_S.

Taking crystal 1 as an example with C_S at 0.01 pF and pulling ability at ± 38.5 ppm, then C_S for the average crystal with ± 60 ppm pulling ability = $(60/38.5) \times 0.01 = 0.016$ pF. We can now also calculate F_S for this crystal using the formula for F_P and transposing: $F_S = F_P/[1 + C_S/2\ C_L]$, which, with $F_P = 10$ MHz and $C_L = 26$ pF = 9.996923 MHz. As this is the crystal we will be working with, and for easy reference, we place the parameters on one line: $F_S = 9.996923$ MHz, $C_S = 0.016$ pF, $F_P = 10.000{,}000$ MHz, $C_L + C_P = 26$ pF.

table 4. Results.

temperature	-2.0 F (-19 C)	25 F (-4 C)	52 F (11 C)	79 F (26 C)	106F (41 C)	133 F (56 C)	160 F (71 C)
crystal 8' (ppm)	21	18	10	0	-10	-17	-21
corrected (ppm)	-22.595	-15.182	-7.634	0	7.869	15.821	23.906
error (ppm)	-1.595	2.818	2.366	0	-2.131	-1.179	2.906

table 5. Results.

temperature	2.0 F (-19 C)	25 F (-4 C)	52 F (11 C)	79 F (26 C)	106 F (41 C)	133 F (56 C)	160 F (71 C)
F_P (MHz)	9.999905	9.999945	9.999985	10.000026	10.000067	10.000108	10.000150
ΔF (ppm)	-9.474	-5.485	-1.460	2.604	6.704	10.842	15.017
C_L (pF)	20.817	20.463	20.115	19.773	19.437	19.107	18.782

table 6. Corrected figures.

temperature	-2.0 F (-19 C)	25 F (-4 C)	52 F (11 C)	79 F (26 C)	106 F (41 C)	133 F (56 C)	160 F (71 C)
ΔF (ppm)	12.078	-8.089	-4.064	0	4.1	8.238	12.413

Next, determine and tabulate the values of C_L, F_P, and the change of F_P (from that at 79 degrees F (26 C) in frequency and ppm with temperature. Except for the frequency, it serves no practical purpose to tabulate beyond three decimal points; similarly, there is little to be gained in making the temperature steps closer than 27 degrees. An operating range of 81 degrees F (45 C) above and below 79 degrees (26 C) is sufficient for most applications.

capacitors with N2200 temperature coefficients

In the following example, C1, C2, and C3 have an N2200 temperature coefficient. And since they are all in series (**fig. 4A**) it is a simple matter to calculate C_L at 79 degrees F (26 C). $C_L = 1/(1/27 + 1/75) = 1/(0.0370 + 0.0133) = 19.853$ pF. Next, calculate the new value of C_L at a temperature 27 degrees F (15 C) below 79 degrees F (26 C). We can use either the formula above or use the calculator program shown in **appendix A**. As this calculation is very simple (there being no mixture of TCs) we will use the formula, thus: $2200 \times 15/10{,}000 = 3.3$ percent. The new capacitive value is $= 19.853 \times 1.033 = 20.508$ pF. That is, for every 27 F (15 C) degree decrease in temperature, the capacitance increases by 3.3 percent. Conversely, the capacitance decreases in value by 3.3 percent for every 27 F (15 C) degree increase in temperature. Hence the multiplier is then $1/1.033 = 0.968$. It is this value that is used in the calculator program in **appendix B** as the capacitance multiplier.

To use the program we enter in **lines 02, 03, 04** the capacitance multiplier 968 plus the other values as shown (which belong to the crystal.) We then start calculations by entering into the calculator display the value of C_L at 15 degrees C below the lowest temperature of interest, thus in the example at -60 C (ΔT), (-29 F or -34 C) $= 1.132 \times 19.853 = 22.47$ pF. *Note that the value of C_L at 79 degrees F (26 C) (19.853 pF) is not used in the program.*

We now have the calculated values of F_P in MHz at every 15-degree C interval as well as the change in F_P (ΔF) in ppm (**table 2**). We find at the 26-degree C point an error of 3.045 ppm. In practice, correction is by adjusting C1 to pull it on frequency. On paper, we do it by substracting 3.045 ppm from each figure (see **table 3**).

Our next step is to look at the curves of **figs. 3** and **5**. These are the frequency/temperature curves for AT-cut crystals. Note the dashed line which intersects the upper and lower turnover points with the various angles of cut. For our purpose, **fig. 5** is more convenient to use. Now project on this set of curves, at the -2.2 F (-19 C) point, the -22.595 ppm obtained from the corrected figures in **table 3**, but with a change of sign. Thus as a + quantity, it intersects the LTP just above the 8 minute cut and in the other direction, at 160 F (72 C), it is close to intersecting the UTP of a 9 minute cut.

A crystal with an LTP at -2.2 F (-19 C) of 22 ppm would appear to be our best bet, (approximately an 8 minute cut). Tabulating the figures for this crystal and subtracting the corrected figures from them, we obtain the results shown in **table 4**.

We have thus improved the frequency/temperature performance of the system from a maximum variation of 42 ppm to one with a change of 5.037 ppm, about an 8:1 improvement.

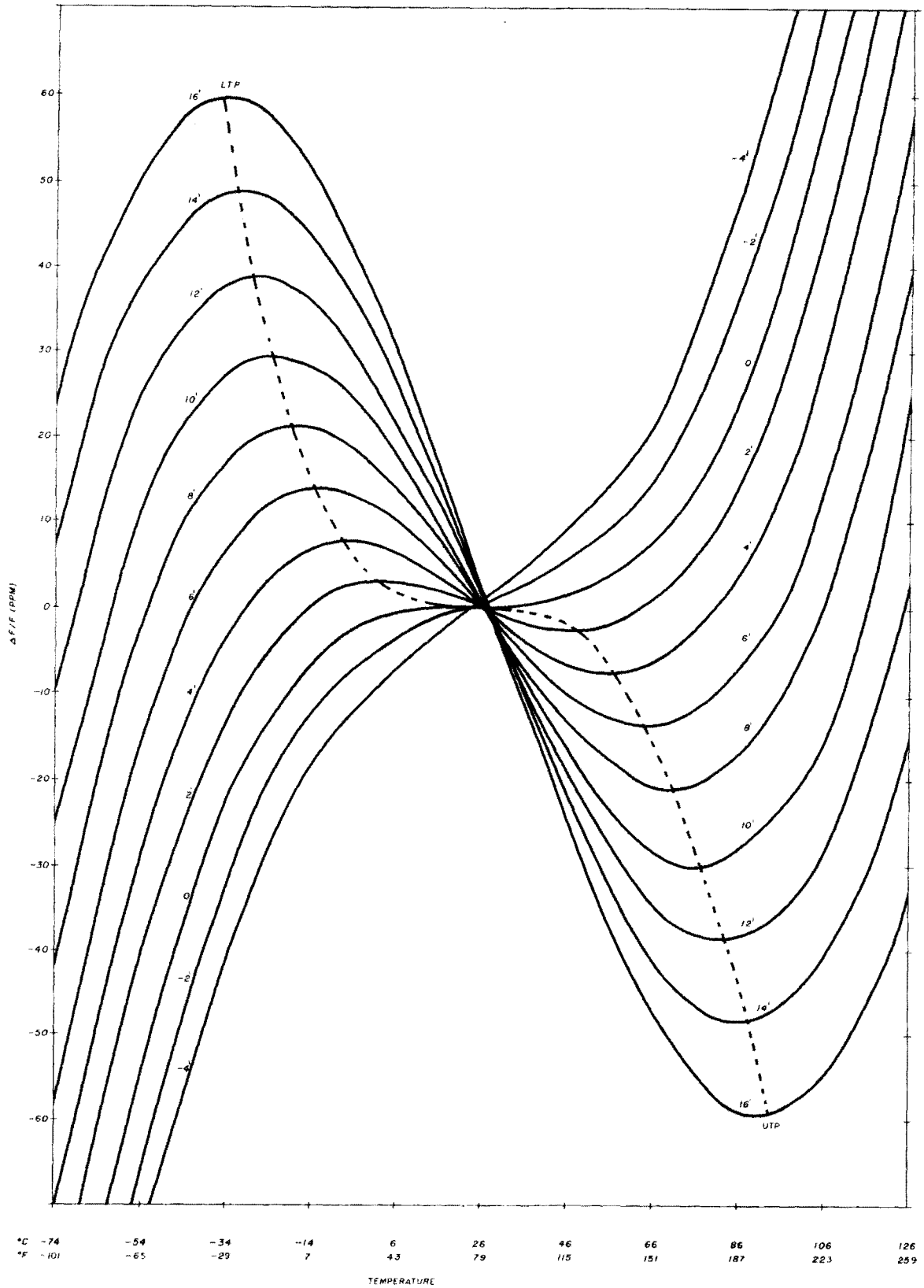


fig. 5. Frequency/temperature curves used to demonstrate crystal performance. (A text example. Chart is for AT-cut crystals.)

table 7. Results.

temperature	- 2.0 F (- 19 C)	25 F (- 4 C)	52 F (11 C)	79 F (26 C)	106 F (41 C)	133 F (56 C)	160 F (71 C)
crystal 5.5′ (ppm)	12	11	7	0	−6.5	−11	−12
corrected (ppm)	−12.078	−8.089	−4.064	0	4.1	8.238	12.413
error (ppm)	−0.078	2.911	+2.938	0	−2.400	−2.762	0.413

N750 temperature coefficient material

In the next example, the TC of C1 is changed to an N750 material. Because we are now employing two different TC materials in the capacitors forming C_L it is easier to make use of the program in **appendix A**. Calculate the value of C_L at 15 C and at −60 C [−29 F (−34 C)] below 79 F (26 C). The rest of the calculations are the same as for our first example. The value of C_L at 79 F (26 C) is of course, still 19.853 pF. Capacitors C2 and C3, being identical and in series, are thus treated as one capacitor of 75 pF.

temperature		52° F (11°C)	−29° F (−34°C)
C1 (pF)	=	27.304	28.215
C2,3 (pF)	=	77.475	84.900
C_L (pF)	=	20.189	21.177

The capacitance multiplier = 20.189/19.853 = 1.017 and the value used in the calculator program is 1/1.017 = 0.983. Start the calculations by entering into the display C_L at −29 F (−34 C) = 21.177 pF.

We have now the calculated values of F_P (in MHz) (see **table 5**) at every 15 C-degree interval, also the change in F_P (ΔF) in ppm. We now correct for the shift at 79 F (26 C), by subtracting 2.604 from each reading see (**table 6**).

Turning now to the curves (**fig. 5**) locate at −2.2 F (−19 C) the +12.078 ppm point. This is located just below the +6 minutes cut, which has its LTP at 7 F (−14 C) and at the 13.5 ppm point. Thus a crystal, having an angle of cut just a fraction below this would probably be our best bet, a cut having its LTP at: 14 F (−10 C) degrees with +12 ppm (approximately a 5.5 minutes cut). Tabulating the figures for this crystal and subtracting the "corrected figures" from them, we obtain the results shown in **table 7**. We have thus improved the frequency/temperature performance from a 24 ppm change to 5.848 ppm, an improvement of 4:1.

summary

The improvement in frequency drift with temperature is considerable and obviously desirable. Instead of specifying the angle of cut, specify either the LTP or the UTP (whichever is more important), giving both the temperature and the amount of change desired from the 79 F (26 C) degree point. Since these points are easily measured using standard test equipment, the manufacturer decides on the angle of cut. Crystals tend to show hysteresis, or inability to retrace reverse conditions, so temperature/frequency measurement should always be made from the same direction in terms of temperature.

To obtain the greatest frequency stability with temperature and time variations, and if power conservation is of secondary importance, place the complete oscillator in a temperature controlled oven, specify the crystal frequency and the UTP at the oven temperature, and allow both to operate continuously.

Where power preservation is important, such as in portable equipment, use NTC capacitors to compensate the crystal and specify the upper and lower turnover points. Place the capacitors close to the crystal so that they also "feel" the same temperature, keeping them clear from heat generating parts.

Overtone crystals are frequently used in the resonant mode. It is generally desirable to tune out Cp with an inductance to prevent oscillations at a frequency determined by this capacitance. Frequency adjustment by the addition of a small amount of inductance is possible.

appendix A

TI-55-II programming

Calculates capacitance with change of temperature. Requires three data memories that contain the numerical values of C, the temperature change or the temperature coefficient, and the number 1,000,000. Displays the new capacitance value.

step	key	line no.	key code	remarks
1	x	00	65	
2	RCL	01	71	
3	1	02	01	memory 1, stores TC or temperature change
4	÷	03	55	
5	RCL	04	71	
6	2	05	02	memory 2, stores 1,000,000
7	=	06	95	
8	×	07	65	
9	RCL	08	71	
10	0	09	00	
11	=	10	95	memory 0, stores the value of C
12	-	11	75	
13	RCL	12	71	
14	0	13	00	
15	=	14	95	memory 0, stores the value of C
16	+	15	94	
17	RST	16	22	displays the new value of C and resets the calculator to start
18	LRN			takes the calculator out of the learning mode.

Now, key in the value of C(27) and STO in 0. Place TC(750) in STO 1 and 1,000,000 in STO 2. Clear display and press RST. Key into the display the temperature −50 C. press R/S. The display should show 28.0125, thus the capacitance increased with a reduction of −50 C in temperature.

Next place into the display 50, then R/S, and the display will show 25.9875, thus a decrease in capacitance with increasing temperature. Thus, by placing the TC of the capacitor into memory 1, and with the capacitor's value stored in memory 0, we can conveniently find the new value with any change of temperature. Interchanging TC with temperature, thus placing the temperature change in memory 1 and keying the TC into the display, will show the effect of changing TC on the capacitance value with various temperature coefficients.

appendix B

TI-55-II programming

Calculates and displays each new value of F_P, ΔF from F_P at 26 C degree, and the new value of C_L (in pF). Note that F is shown in + or − ppm. Requires three data memories. Memory 0 stores the new (decremented) value of C_L, memory 1 contains F_S in MHz (to six decimal places), and memory 2 contains the value of F_P at 26 degrees C also to six decimal places.

step	key	line no.	key code	remarks	example
1	×	00	65		
2	.	01	93		
3	'	02	'		9
4	'	03	'	enter: capacitive multiplier	6
5	'	04	'		8
6	=	05	95		
7	STO	06	61	memory 0 must be empty (0)	
8	0	07	00	at the start	
9	+	08	85		
10	'	09	'	value of C_P (6 pF)	6
11	=	10	95		
12	×	11	65		
13	2	12	02		
14	=	13	95		
15	÷	14	55		
16	'	15	93		
17	'	16	'	value of C_S (0.016 pF)	0
18	'	17	'		1
19	'	18	'		6
20	x-y	19	52		
21	+	20	85		
22	1	21	01		
23	=	22	95		
24	×	23	65		
25	RCL	24	71	enter F_S in memory 1	9.996923
26	1	25	01		
27	=	26	95		
28	R/S	27	12		
29	.	28	75		
30	RCL	29	71	enter F_P at 26 degrees C	10.000000
31	2	30	02	in memory 2	
32	−	31	95		
33	÷	32	55		
34	RCL	33	71		
35	2	34	02		
36	=	35	95		
37	2nd Eng	36	47		
38	R/S	37	12		
39	RCL	38	71		
40	0	39	00		

End of program. Automatically resets and displays the new value of C_L, which is now in memory 0.

To start, press RST, enter F_S in memory 1 and F_P in memory 2. Check that memory 0 is 0. Enter into the display the value of C_L at −76 F (−60 C). Example 22.474. Press R/S. Displays F_P at 2 F (−19 C) = 9.999804. Press R/S, display −19.5507 06. This is the change in frequency from F_P at 26 C = −19.550 ppm. Press R/S, display = 21.754832 00; thus 21.755 pF, which is C_L at −2 F (−19 C). Press R/S again. This repeats the calculations, using this time C_L at −2 F (−19 C), calculates all the 25 F (−4 C) parameters, and so on, saving a lot of repeating calculations.

appendix C

TI-55-II programming

To calculate $F_P = F_S\ [1 + (C_S/2C_L)]$ requires four data memories that will contain the numerical values of C_S, C_L, F_S, and F_P at 26 C degrees in this order. In the same order will be displayed the values of $C_S/2C_L$, F_P and the difference between F_P and the F_P at 79 F (26 C), thus ΔF. Partition four memories and place the calculator in the learn mode.

step	key	line no.	key code	remarks
1	RCL	00	71	
2	0	01	00	memory 0 stores C_S
3	÷	02	55	
4	(	03	53	
5	2	04	02	
6	×	05	65	
7	RCL	06	71	
8	1	07	01	memory 1 stores C_L
9	)	08	54	
10	−	09	95	
11	R/S	10	12	displays value of $Cs/2C_L$
12	+	11	85	
13	1	12	01	
14	−	13	95	
15	×	14	65	
16	RCL	15	71	
17	2	16	02	memory 2 stores the value of F_S in MHz
18	−	17	95	
19	R/S	18	12	displays the value of F_P in MHz
20		19	75	
21	RCL	20	71	
22	3	21	03	memory 3 stores F_P at 26 C degree point, (thus with a total C_L at 26 pF). Example 10.000000 MHz)
23	.	22	95	
24	RST	23	22	displays the difference between F_P and the value contained in memory 3
25	LRN			takes calculator out of learning mode.

Key in the value of Cs (0.01) and STO in 0. Place C_L (26) in STO 1, F_S in STO 2, and in STO 3. Place 10,000,000. Now press ON/C to clear the display. Check the contents of each memory for errors. If OK again clear the display and press R/S to run the program. The first stop should be $Cs/2C_L$ = 0.0001923. Press R/S again, up comes F_P = 10.000166 and finally the frequency difference at 0.0001667 MHz. Press 2nd, Fix 6, which will fix the display to 6 decimal points.

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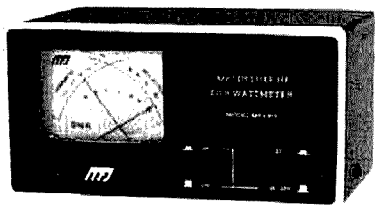
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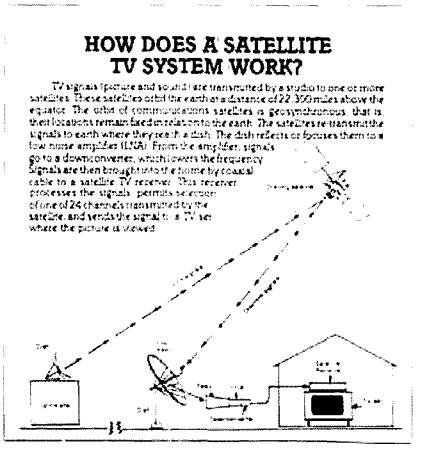


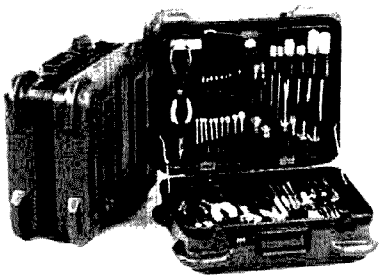
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How To Read What You Writ! is available directly from Loraine McCarthy, Code Programs Manager, Gordon West's Radio School, 315½ Ruby, Balboa Island, California 92662. The price is $14.95. All orders are shipped by first class mail the same day. (Add $2.00 postage and handling.)

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Heathkit catalog

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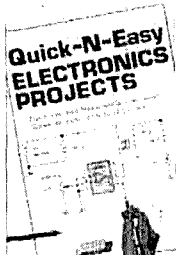
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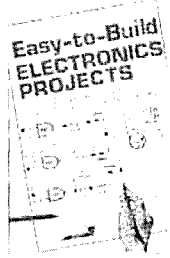


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COMING EVENTS

Activities — "Places to go . . ."

CALIFORNIA: FCC exams, Novice Extra. Sunnyvale VEC ARC (408) 255-9000 24 hour. 73, Gordon, W6NLG, VEC.

PENNSYLVANIA: The Mercer County Amateur Radio Club invites Amateurs to socialize and exchange views at its "How To" seminar, March 22, 9 AM to 4:30 PM, Hermitage Middle School opposite Shenango Valley Mall, Rt. 18, Hermitage. License exams. Novice through Extra, will be given. Examinees and those under 18 years admitted free. For others admission is $2.00. Call in on 147.75/.15. For details SASE to MCARC, Box 996, Sharon, PA 16146.

MASSACHUSETTS: The Mt. Tom Amateur Repeater Association is having its annual Flea Market, March 2, Knights of Columbus Elder Council 69 on Granby Road, Chicopee. Doors open 8 AM to 3 PM. Food available. Reserved tables $7.00. At the door $8.00. Entrance fee $1.00. Ladies and children under 12 free. For reservations: MTARA, PO Box 3494, Springfield, MA 01101.

MINNESOTA: The Robbinsdale Amateur Radio Club's 5th annual Midwinter Madness Hobby Electronics Show, February 22, Totino-Grace High School, 1350 Gardena Avenue NE, Fridley, MN (suburb of Minneapolis). Admission $3.00 advance; $4.00 at the door. 8' flea market tables $8.00, 1/2 tables $4.00. Dealers, flea market, computers, satellite TV, packet radio. Talk in on 147.60/.00 K0LTC repeater and 146.52 simplex. To register send SASE with fees to Robbinsdale ARC, PO Box 22613, Robbinsdale, MN 55422 or call Bob (612) 533-7354. FCC exam registration send completed Form 610, photocopies of current license and code credit plus $4.00 payable to ARRL/VEC to: Neil McMillin, 11132 97th Place, Maple Grove, MN 55369 by January 22, 1986. Limited walk-ins.

KENTUCKY: Annual Glasgow Swapfest, Saturday, February 22, 8 AM till Glasgow Flea Market Building, 2 miles south of Glasgow off Highway 31E. Free parking. Free coffee and a friendly gathering of hams. Admission $2. No extra charge for exhibitors. One free table per exhibitor. Extra tables $3.00 each. Talk in on 146.34/94. For information: N4HCO, Rt 4, Box 354, Glasgow, KY 42141.

MASSACHUSETTS: The Norwood ARC will hold its annual Flea Market, Saturday, February 22, Norwood Junior High School South, Washington Street. Dealer tables $10.00. Setup 8 AM. General admission $2.00. Open 9 AM. Plenty of free parking and facilities for the handicapped. For information or table reservations: Stan Cottrell, WA1NCV (617) 762-5184. Talk in on 146.520 and 146.895.

ILLINOIS: The Sterling-Rock Falls Amateur Radio Society's 26th annual Hamfest, March 9, Sterling High School Fieldhouse, 1608 Fourth Avenue, Sterling. Distributors, dealers, large flea market and demonstration of Packet Radio. Free parking and space for self contained RV's. Doors open 7:30 AM. Tickets $3.00 advance; $4.00 at the door. Commercial tables $5.00. Others $3.00. For tickets, tables or information: Sue Peters, KA0GMR, PO Box 521, Sterling, IL 61081, (815) 625-9262. Talk in W9MEP 146.25/.85.

MASSACHUSETTS: The MIT UHF Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday, February 19, 1986, 7 PM, MIT Room 1-134, 77 Mass Ave, Cambridge, MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-5820/646-1641 or Craige Rodgers at 225-6616. Exam fee $4.00. Bring copy of current license, 2 forms of picture ID and completed form 610 (available from FCC in Boston. 223-6609.

VIRGINIA: The Vienna Wireless Society will hold its 13th annual Winterfest, February 23, 8 AM, Vienna Community Center, 120 Cherry St., Vienna. Exhibits and demonstrations of new and used Amateur Radio equipment. Tailgating in the "Frostbite" parking lot. Refreshments available. For information: John Arnold, N4IXD, Chairman, 255-2076.

Operating Events — "Things to do . . ."

North Carolina QSO Party sponsored by the Alamance ARC, K4EG, 1400Z February 1 to 0500Z February 2. CW and phone. Work stations once per band/mode. Mobiles as they change county. Exchange RS(T) and ARRL section. Mail logs by 3/1/86 to North Carolina QSO Party, c/o K4EG, PO Box 3064, Burlington, NC 27215.

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THE GUERRI REPORT

by Ernie Guerri, W6MGI

millimeter waves

Part I: the move to higher frequencies

For most Amateurs, the notion of moving to higher frequency bands means operating at or above 144 MHz. A few brave pioneers work in the microwave region — and have precious few contacts to show for it. We could be tempted to forget that for more than 40 years, there has been an avalanche of effort devoted to the development of microwave hardware and systems.

For the purposes of this discussion, "microwaves" will mean the region above 2 GHz — that is, up to about 30 GHz. This is the domain where lumped circuits fail to have physical implementation and where waveguide replaces coax in many low-loss or high-power applications. While military uses were the principal motivator for early work, commercial applications are now the driving force in the region from 2 to 18 GHz.

Band crowding is already a major issue in the microwave region. By the mid-1960s, the military had found use for nearly everything up to about 20 GHz. Then came the demands for commercial satellite uses, followed by the expansion of terrestrial links to take advantage of all the information the satellites were sending. By the early 1980s, interference was an issue in some frequency bands. The result has been a critical examination of the technologies needed to make higher frequencies available for both commercial and military applications.

These higher frequencies are known as *mm waves* — and, for convenience, are generally defined as the wavelengths from 10 mm to 1 mm (30 to 300 GHz). One of the obvious advantages of this new domain is the amount of spectrum space available: 270 GHz.

Among the advantages that mm systems offer are broad bandwidths (good for high speed data links), higher spatial resolution, less interference due to narrow antenna beams, and, at the high end of the range (near 300GHz), a favorable alternative to optical (infra-red) technologies.

Much work has already been done in this region, but the advent of computer modeling techniques is advancing the rate at which progress can be made. A major limitation on the use of mm waves has been the lack of appropriate devices, hardware, and test equipment. Within the past two years, however, there have been significant developments in each of these areas.

The generation and detection of mm waves requires the use of components and techniques that are unfamiliar to most Amateurs. Semiconductor devices are the only sources of RF power at frequencies above 30 GHz. In general, these devices take the form of negative resistance oscillators and depend on the characteristics of special diodes. The more common of these diodes are GUNN diodes, named for the discoverer of transferred electron devices; IMPATTs (*I*mpact *I*onization *A*valanche *T*ransit *T*ime); and TRAPATTs (*T*rapped *P*lasma *A*valanche *T*riggered *T*ransit).

Each of these devices has characteristics that make it best for certain applications. The tradeoffs are between frequency (of operation), power output, and noise. IMPATT dioides have been fabricated that can produce 1 watt at about 70 GHz.

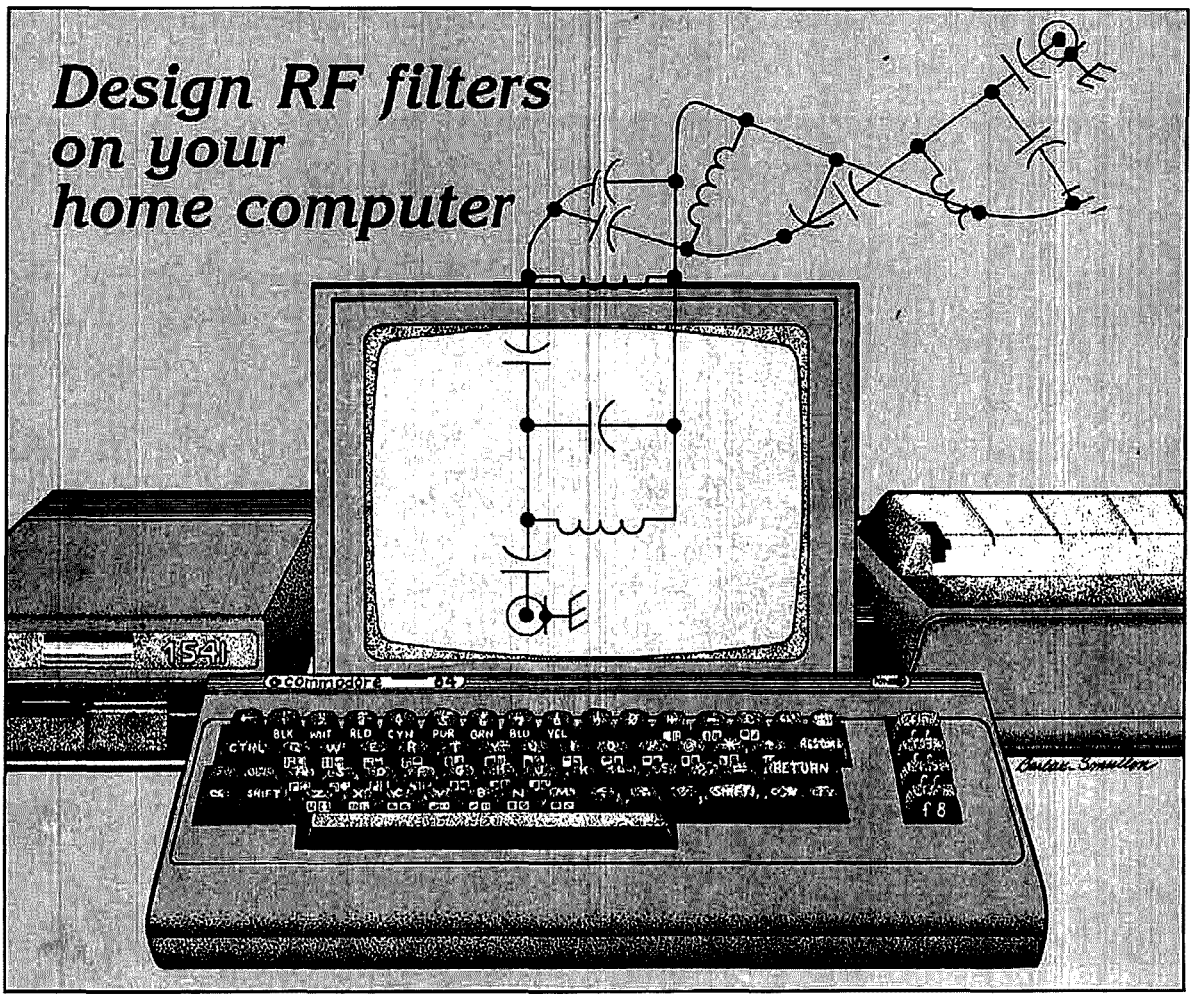
Transistors will soon be available for the mm region. GaAs FETs are already being made as small signal amplifiers above 30 GHz, and it shouldn't be too long before HEMT (*H*igh *E*lectron *M*obility) and MESFET (*M*etal *o*n *O*xide) devices are available to 100 GHz or so. Power transistors for this region are still a few years off — but only a few. Other necessary components are being developed quite aggressively, as is test equipment. Hewlett Packard has just announced its first line of mm wave spectrum/network analyzers. Special materials and design techniques will be needed to take best advantage of the promise held by mm waves. Work is underway in each of these areas, and the results will not be more than four to five years away.

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Next month: Part II — Where Do We Go From Here?

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Michael Gruchalla, P.E.
Bob Lewis, W2EBS
Mason Logan, K4MT
Ed Wetherhold, W3NQN

publishing staff

J. Craig Clark, Jr., N1ACH
assistant publisher

Rally Dennis, KA1JWF
director of advertising sales

Dorothy Sargent, KA1ZK
advertising production manager

Susan Shorrock
circulation manager

Therese Bourgault
circulation

cover art:
Barbara Smullen

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REFLECTIONS

an increasingly important design tool

Years back the typical approach to filter design was to decide on the required characteristics, choose the best topography, or type, consult charts (e.g. the ITT *Handbook*), evaluate formulas, and — voila! — the circuit evolved.

Perhaps after this procedure was applied it met one's specific goals in terms of passband, stopband, ripple, and VSWR responses, in addition to input and output matching. At best, however, even at this paper design stage, it involved an iterative process whose speed of convergence to the required performance parameters greatly depended on the accuracy of the educated guesses made by the filter designer. A designer with good hands-on RF experience was often successful, but nevertheless needed time to determine if the design was truly optimum.

The fun would then begin in earnest with the selection of physically-realizable components (reasonable inductor *Q*), construction, and tweaking. If the designer was fortunate enough to have a sweeper, delay lines, a wide video bandwidth scope, and other necessary test instrumentation, an actual circuit could be built, aligned, and applied. But even then the nagging thought would persist: *is* this an optimum design? And what if one particular characteristic — ripple, perhaps — were changed for improvement in another response . . . better stopband rejection, for example?

This process changed, to a large extent, with the advent of the personal computer and filter software routines. Thanks to programs like the one WB4EHS describes in "Build Narrowband RF Filters," this month's lead article, it's possible for a newcomer to filter design to input his requirements and let the microcomputer manipulate the formulas and other data, thereby effecting the best possible design in *seconds*. If this weren't enough of an improvement, the designer can now interact with the computer and investigate other possibilities if further tradeoffs are to be considered.

At *ham radio* we recognize the importance of the use of computers as useful tools for the Radio Amateur and have included, in past issues, articles describing the use of computers and appropriate software to aid in the design process, much as slide rules and hand-held calculators were used before. Review the five-year cumulative index in the December, 1985, issue for reference to recent computer-related design articles.)

This doesn't mean what we're becoming yet another computer-oriented magazine . . . but on the other hand, we won't ignore relatively new procedures that provide useful circuits with greater speed, accuracy, and applications.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

A COURT RULING OF GREAT POTENTIAL POSITIVE BENEFIT TO AMATEUR RADIO has been reached in the case of Thernes vs City of Lakeside Park (Kentucky). In its decision in favor of John Thernes, WM4T, in his three-year-old antenna dispute with Lakeside Park, the U.S. Court of Appeals for the 6th Circuit ordered an earlier district court's judgment against Thernes vacated and the case remanded to the district court for reconsideration.

In A "Dissenting" Opinion Judge Krupanski Argued The Two Other Judges weren't hard enough on the city; referring to PRB-1, he said the city's anti-Amateur antenna ordinance was "arbitrary and capricious and deprived the appellate of equal protection under the law" since it applied to Amateur antennas but not similar TV structures. He condemned the city's position as "patently frivolous, discriminatory, and without factual support" and the ordinance as "a license to discriminate," concluding "I would reverse the (lower court's) decision, declare the ordinance unconstitutional, and award costs against the city including attorney's fees."

Strong Language, Indeed, In This First Application of PRB-1, even though the issue in W4MT's case is far from resolved. Now it's up to the original trial court to make the determination of the effect of the FCC's preemption on the constitutionality of Lakeside Park's antenna zoning ordinance. If that decision favors Amateur Radio (assuming Lakeside Park even wishes to continue the long-term court battle), then PRB-1 should be a very strong influence in favor of Amateur antennas and towers throughout the country.

The FCC Has Also Adopted A Limited Preemption Ruling On Satellite Dishes vs zoning and other local ordinances. In a January 14 decision, the Commission said, in essence, that local regulation should apply equally to all types of antennas, and satellite earth station antennas must not be burdened with special limitations not applicable to other antennas.

"AUTOMATIC CONTROL" FOR DIGITAL AMATEUR COMMUNICATIONS ABOVE 50 MHz was approved by the commission in a January 13 report and order on PR Docket 85-105. However, there will still be an important limitation with respect to packet since Part 97 requires a control operator be present when third-party traffic is being handled and defines third-party traffic as any message not originated by the station operator.

EXPANSION OF 40-METER PHONE FOR U.S. CARIBBEAN AREA AMATEURS has been announced by the FCC; Generals, Advanced, and Extras may operate phone from 7075-7100 kHz from Puerto Rico, Virgin Islands, and other FCC-administered Caribbean areas after 0001Z February 28.

THE AMATEUR MILLIMETER WAVE SPECTRUM WILL SEE MAJOR CHANGES ON MARCH 1, with the loss of 260 MHz plus the 1215-1240 MHz slice of the 23 cm band. No immediate crowding of the Amateur mm-wave bands is likely, however, as we'll still have over 12,000 MHz of mm-wave spectrum available. The final implementation of WARC '79 is the reason; the new mm-wave frequencies are 47.0-47.2, 76-81, 119.98-120.02, and 142-149 GHz. the deleted bands are 48-50, 71.0-71.5, 165-170, and 240-241 GHz.

ALL FOUR "INDEPENDENT" NATIONAL VECS HAVE NOW JOINED CARE, the national organization established to support and promote the Volunteer Examination program. In addition several regional VECs have also signed up, along with a number of individual VEs including some accredited by the ARRL. Inquiries about CARE can go to Jim Georgias, W9JUG, at DeVry Institute of Technology, (312) 929-8500, or to CARE, Box 688, Glenview, IL 60025.

"Zero Defects" For Their Paperwork Was Achieved By All VECs for the first time during the month of December! Paperwork errors have been a major VEC program problem since its inception, but continued pressure from the FCC and educational efforts by individual VECs finally paid off when not one error was found in 3651 sets of December exam submissions.

Tardy Submissions Have Also Dropped Drastically over the past year; in December, 1984 27 percent arrived late at Gettysburg. Just under 3 percent were past deadline last December.

POSSIBILITY OF A NEW "CONSUMER RADIO SERVICE" WILL BE THE TOPIC of an FCC meeting scheduled as this Presstop goes to press. The commission is expected, in response to the first of several petitions by the Personal Radio Steering Group, to inaugurate a study of the public's need for radio communications and how it relates to the General Mobile Radio Service, the sophisticated UHF portion of the Citizens Radio Service. The PRSG has proposed an evolutionary change of GMRS into a more flexible service utilizing narrow-band technology and directed toward personal rather than business needs.

A SIZEABLE INCREASE IN NOVICE LICENSES LED THE U.S. AMATEUR POPULATION to another all-time high of 415,856 at the end of December. All license classes showed increases over the preceding month, but Novices were up a whopping 1529! According to FCC analysis, 2233 Novices were newly licensed in December, 1985, versus only 818 the month before and 1196 in October. In addition the December, 1984, figure was only 1409 new Novices, so there's apparently a highly successful upsurge in Novice training efforts. ARRL has also launched a strong effort to get Novices who've let their licenses expire in the past two years renew, with expectation that a valid license in hand will restimulate interest.

With the VEC Program Running Smoothly And Recruitment/Training paying off so well, 1986 could be a banner year for the Amateur Service even without new Novice privileges.



comments

digital can do more

Dear HR:

I want to let you know how much I enjoyed the December issue of *ham radio*, with its emphasis on Digital Communications. I think you should be commended for your efforts to move Ameteur Radio into the digital age. There are, however, a few comments I wish to make on the subject.

First, we must get across the idea that "digital" and "data" are not synonymous. There are many uses for digital transmission besides that of data. For example, digital voice, digital graphics, and digital video (both freeze frame and full motion) are areas where the Amateur community should be conducting experiments and demonstrations. Packet radio for data is great, but so much more can be done. Very sophisticated digital voice processors should soon drop in price (some, such as CVSD chips, are cheap now) and hopefully, 2.4 kbps LPC voice processing chips such as the Texas Instruments 320.

Second, I believe the data rate limitations you quoted in your editorial on page 5 (and also quoted on page 30 [by WA1FHB in "A Packet Radio Primer" — Ed.]) are too conservative. Paragraph (a)(1), Part 97.69 of the FCC Regulations on Digital Communications states that "The sending speed shall not exceed the following:" and goes on to give the limit in baud depending on the frequency of transmission. The definition of a baud is a unit of modulation rate. Modern digital radio systems may have many bits per second for each baud. Four-level, eight-level, 16-level, 64-level and 256-level systems have been demonstrated and many of these are used in microwave systems today. Our radio, the International Mobile Machine Corporation's *Ultraphone*™ has demonstrated 64 kbps with 16-level modulation in an RF channel not much wider than a 20 kilohertz (4 bits per signaling element). This was done on a frequency of 455 MHz. With the same 4 bits per signaling element, this would allow 1200 bps on frequencies below 28 MHz, 4800 bps on frequencies between 28 and 50 MHz, 78.4 kbps on frequencies between 50 and 220 MHz and 224 kbps above 220 MHz — a 4:1 expansion of your limits. It should be noted that you don't get something for nothing; these high level modulation schemes require a higher signal-to-noise ratio and more signal processing.

It is well known that in some cases, combining high level modulation with the proper error correcting code can enhance the performance of the link over lower level modulation. The article on Hermes ["AMTOR, AX.25, and HERMES," by Jerome T. Dijak, W9JD, — Ed.] was a good start on the right type of coding. Much work needs to be done on both modulation and coding before standards are set. However, it would be nice to see someone coming out with PSK/DPSK radios.

To give one example, Bolt Beranek and Newman, Inc. have been working on digital voice at 300 bps (see *Electronic Design*, October 3, 1985, page 92). Although still experimental, and requiring a ridiculous amount of computing power (256 68000 processors), this, when combined with a four-level DPSK modulation with a 1/2 rate coding (and interleaving) would give a very interesting HF system running at 300 baud. A good experiment would be to determine the amount of interleaving processing delay that would be tolerable.

A second near term example might be using 2.4 LPC voice processing with a four-level dpsk 4800 bps modem and 1/2 rate error correcting coding on 10 meters. This system should be feasible for Amateurs to construct in the next year or so at a reasonable cost.

Another example might be high resolution (i.e., 600 pixels) freeze frame video (or computer-generated graphics or fax) with errorless packet transmission at 4800 baud on 10 meters. This might take 10 to 15 minutes to send, but would provide a high quality, error-free picture. Signal processing of the image could lower the transmission time significantly. New, lower-cost CCD video cameras will make this concept very feasible.

A fourth example might be full-motion, color video on 6 and 2 meters and higher. Systems now exist that can transmit these at 56 kbps and with VHSIC technology, the cost could come down to make this feasible for Amateurs.

Some of the above examples are still far out, but still are reasonable goals for Amateur Radio. I think that the future is digital radio and that low cost, spectrally efficient modulation and coding are the keys that will make digital radio feasible for Amateurs. Fortunately, this modulation and coding can be implemented in software on signal processing chips such as the Texas Instruments TMS-320. We need two things to happen: one is for knowledgeable Amateurs to develop these new modems, and the second is for manufacturers to develop radios that will accept these modems.

Allen D. Dayton, KA4JFO
Fairfax, Virginia

build narrowband RF filters

...and learn useful RF techniques as well

LC bandpass filters have been extensively covered in the professional literature. While some of this has spilled over into the Amateur press, some confusion still remains about what type of filter to use in a given application, and how to go about designing, building, and aligning it. In this article I hope to clear up some of this confusion and help you choose, design, build, and evaluate the filter you need. The BASIC program (written in Commodore 64 BASIC, but translatable to others) will be presented to perform the math for you.

The filters we're describing here are narrowband, capacitively-coupled designs (see **fig. 1**; note that the program prints out the capacitor values using these designators, so keep this figure in mind for future reference.) By narrowband, I'm specifically referring to the results of the following, well-known relationship[1,2,3]:

$$Q_{bp} = \frac{f_o}{BW_{3dB}} \quad (1)$$

f_o = mean center frequency (below)
BW_{3dB} = passband width

For these filters, Q_{bp} must be ≥ 10. For $Q_{bp} < 10$ other design techniques are better. References 1 and 2 are recommended for coverage of this sort of design, based on lowpass filters, and for converting them to bandpass.

Like most bandpass filters, these filters have a geometric symmetry about the center frequency:

$$f_o = \sqrt{f_u f_l} \quad (2)$$

f_u = upper passband edge
f_l = lower passband edge

Because of this, the response of the attenuation versus frequency of these filters does not follow the arithmetical symmetry one might expect. Attenuation increases more rapidly on the low side of the passband.

Also, like most BPFs, these filters have passband versus stopband characteristics that are inevitably a compromise between passband ripple and attenuation at some out-of-band frequency. The optimally flat Butterworth (no-ripple) designs will have the poorest attenuation out-of-band. The Chebychevs, with varying amounts of ripple, will give more attenuation at the same out-of-band frequency, with the highest ripple designs giving the largest amount of attenuation. This attenuation also carries the price of being the most difficult of the filter types to build with real-world components.

Choosing the right filter for a job, then, becomes an exercise in compromise. In the front end of a receiver, for example, the ripple of the Chebychev might mean stations not copied (DX not worked). You might put up with a little less rejection of an out-of-band signal to hear those. Later on in the rig, a little gain can frequently be spared — perhaps the Chebychev should go there. One of the advantages of the program is that it allows you to compare designs before you plug in your soldering iron.

By Bob Lombardi, WB4EHS, 1874 Palmer Drive, Melbourne, Florida 32935

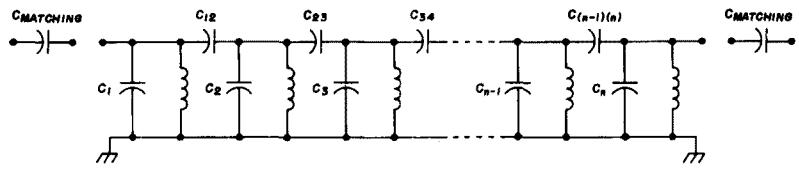


fig. 1. General form of filter, showing matching capacitors and their placement.

using the program

To design a filter with this program, you need to know a few things: first, the center frequency (f_o) and passband width. Both of these are entered in MHz. Next you need to know the required passband characteristics (i.e., Butterworth or 0.01 dB, 0.1, 0.5., or 1.0 dB ripple Chebychev), what inductor you plan to use, and how much attenuation is required at some frequency ratio.

This last one may be unfamiliar to you. This frequency ratio is expressed as:

$$\frac{f}{f_c} = \frac{BW_x}{BW_3} \qquad (3)$$

where BW_x is the bandwidth at some attenuation x, and BW_3 is the (3dB) passband width, as defined before. Often, you won't know BW_x directly, but will know that there is some undesired frequency (f_{un}), such as a broadcaster, or mixing product that you want to attenuate. You can solve for BW_x by using the geometric symmetry of the filter, as follows:

$$f_o = \sqrt{f_{un} f_{ot}} \qquad (4)$$

then

$$\frac{f_o^2}{f_{un}} = f_{ot}$$

and $BW_x = |f_{un} - f_{ot}|$

f_{ot} = the other frequency you need

You then divide this number by the bandwidth to get the ratio, f/f_c, that the progam queries for. It is entered as a single number (i.e., 3 or 4.237) rather than as a ratio.

The filter that results will inevitably have some characteristic impedance that will never be the one you're looking for — per Murphy. To overcome this, a simple matching capacitor can be calculated for you by the program. It can match either end to another impedance, and does this by making the end sections a capacitive divider, with one capacitor in series with the input and output. It will recalculate the changed capacitor in the *body* of the filter for you. The only hazard here is that a large change in impedance may make one of the capacitors go negative in value. If this happens, you can redesign the filter with a bigger inductor, or else use another matching method of your own.

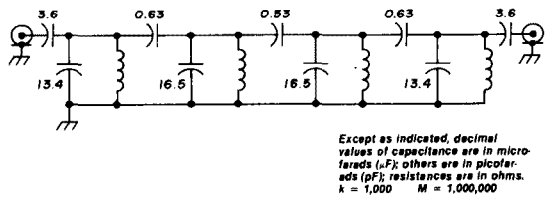


fig. 2. The two meter filter uses 68nH, Coilcraft T-113 1½ turn coils.

sample filter

To illustrate the process, here's a filter that I needed for a 2-meter transverter. The bandwidth at 50 dB down was determined by a local NOAA Weatheradio signal at 162.55 MHz that I wanted to knock down as much as possible. The 3 dB bandwidth of the filter was determined by a need to cover all 4 MHz of 2 meters in the peak of the passband. My inputs to the computer are the requirements that I have for the filter:

Type of filter: 0.5 dB ripple Chebychev
Frequency Ratio f/f_c: 4.13
Desired Attenuation at f/f_c: 50dB

At this point, the computer responds that an N = 4 (4 pole) filter will provide 61.08 dB (at f_{un}) and then asks for the center frequency (145), bandwidth (8), and the coil used (0.068 μH) It then calculates its responses. Its outputs are:

Coupling Capacitors:
C12 = .6334 pF
C23 = .5327 pF
C34 = .6334 pF
Resonating Capacitors:
C1 = 17.08 pF
C2 = 16.55 pF
C3 = 16.55 pF
C4 = 17.08 pF

It concludes by telling me that the characteristic impedance of the filter is 2050 ohms and asks me if I want to match to another. I request 50 ohms, both ends, add 3.47 pF to the end sections and change the end resonators, C1 and C4, to 13.613 pF (17.08-3.47).

Although this filter was touchy to align, it was built and aligned using sophisticated equipment, then measured on a Hewlett Packard Network Analyzer. The graphs of attenuation versus frequency for it are in fig. 3 along with its schematic fig. 2.

building the filters

Of course all this is pretty meaningless without working filters that fit your needs — and the program, like all mathematical models, is perfectly capable of generating answers that solve its equations, but are completely absurd in the real world ... all of which requires the introduction of some sound ideas for building these filters.

To begin with, filters using conventional discrete coils and caps get progressively harder to build and align as frequency increases. The Amateur with the typical test equipment described below and good components and assembly techniques should have no trouble getting these filters to work up through 75 to 100 MHz. Those with high-quality equipment or access to it can expect to go to beyond 2 meters. Above 200 MHz, filter construction and alignment without large amounts of equipment demands other techniques, which will not be discussed here.

When it comes to components, the rule is to get the highest Q that you can obtain. Low Q causes losses in the filter, evidenced by higher insertion loss and changes in the passband shape. Capacitors are rarely a problem because they tend to have much higher Qs than coils. In an inductor:

$$Q = \frac{X_L}{R} = \frac{2\pi fL}{R} \qquad (5)$$

where R is the ohmic resistance of the wire and X_L is the inductive reactance. From this definition, it follows that you should use the largest practical size wire you can. Q is also affected by coil diameter, winding pitch, and core or support material. Adding a core always degrades Q as does shielding a coil.

The tradeoff here is that airwound, large diameter coils are fine at VHF, but too large at HF. Thankfully, iron powder and ferrite toroids provide increased inductance, virtually perfect shielding, and reasonable Qs from MF to the low VHF region. Amidon Associates provides data that allows you to predict Q and the number of turns to wind a given coil, on request.

While Q may not be a major concern with capacitors, stray reactances are. At RF, they contain enough

table 1.

Type of Filter	Minimum Component Q						
n =	2	3	4	5	6	7	8
Butterworth	2	2	3	3	4	4	5
0.10 dB Chebychev	2	3	5	7	10	14	17
0.50 dB Chebychev	3	4	7	10	14	20	30
1.0 dB Chebychev	3	5	8	12	17	26	35

This table gives an approximate minimum component Q for use in a *lowpass* filter of the given order. For a bandpass filter, this value must be multiplied by the passband Q, Q_{bp}.

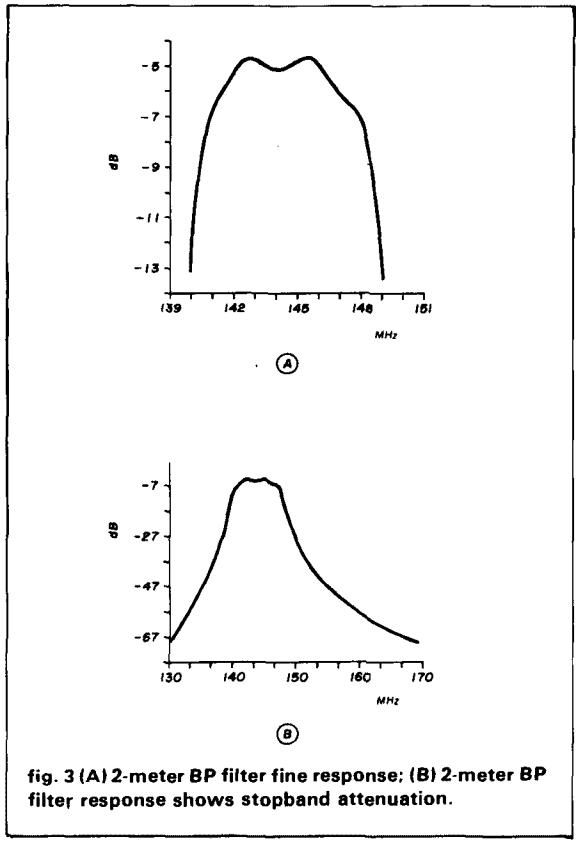


fig. 3 (A) 2-meter BP filter fine response; (B) 2-meter BP filter response shows stopband attenuation.

stray inductance to present themselves as something far from a perfect capacitor. For example, a 0.001 μF ceramic disk capacitor with 1/4 inch leads appears as a series-resonant circuit (0 ohms) at 55 MHz, while a 0.01 is self-resonant at 15 MHz.[4] For this reason, it is imperative to keep leads as short as possible.

As for the type of capacitor to use, ultraminiature dipped ceramics are best at and above 6 meters, although their typical 50 VDC breakdown is low for transmitters. Through the HF spectrum, silver micas are an acceptable substitute. SMs have a higher stray inductance, but it's rarely a problem. They're available in higher voltage ratings, too.

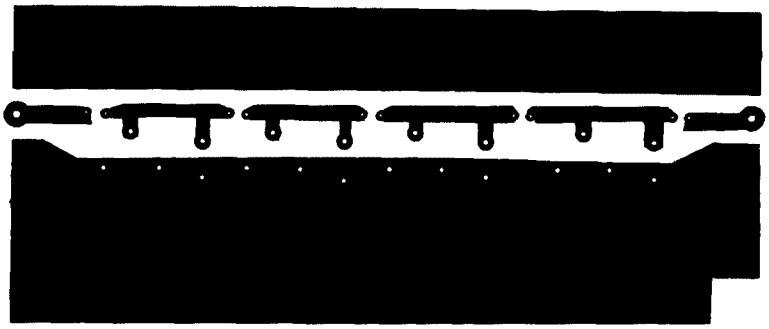


fig. 4. Full size version of the "universal" 4 section filter card.

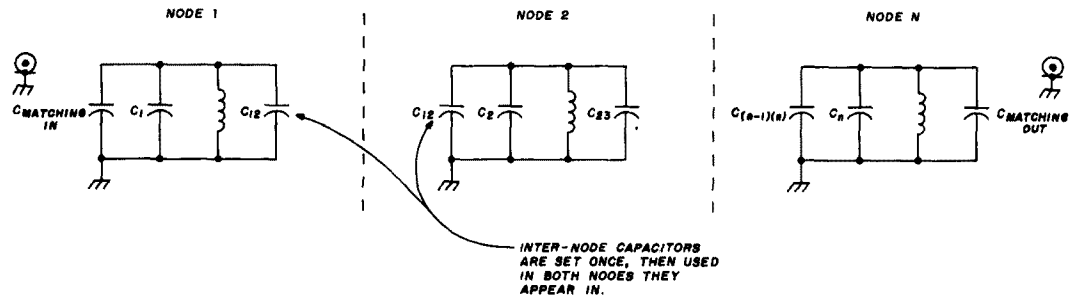


fig. 5. Breaking up a filter into nodes for alignment.

Inductors also show stray reactance — in this case, capacitance between adjacent turns. This has the effect of making the inductor appear smaller than it is, and is especially noticeable as frequency increases.

One last word on component Q. The filter type chosen imposes some constraints on the minimum Q you can live with. **Table 1**, derived from reference 2, presents the approximate minimum Q for a lowpass filter of the given type and order. For bandpass filters, this must be multiplied by the Q_{bp}, derived previously. From this, you can see how high-order Chebychevs would be difficult to realize, based on component Q alone.

Component layout requires some thought. While an etched PC board is not necessary, at least through 6 meters, some readers may want to etch one anyway. The optimum layout is linear, with the components appearing as they are drawn on the schematic. If you must "bend" this line to fit into a space, try to keep input away from output and break grounds between them. A U-shaped layout isn't *asking* for trouble— it's *begging*.

Figure 4 is a "universal" four-pole filter prototyping board I've used from 1 to 200 MHz. It's etched on double-sided G-10, with the back solid groundplane. Components are soldered directly to the lands or the lands are jumpered out with copper foil for smaller numbers of sections. Top and bottom groundplanes are jumpered together by wires.

If you choose not to etch a board, or can't etch a board, take heart. Many prototypes in engineering labs are built in a style called "dead bug"; components going to ground are soldered to a piece of unetched PC material and those not grounded are supported by their leads going to those that are, or to each other. The only precautions necessary are to keep ground connections as short as possible and to not lay unshielded coils with their axes parallel to each other, or with turns touching. Toroids can be placed in any orientation to each other and can touch ground without trouble. As in etched boards, the best layout is a straight line.

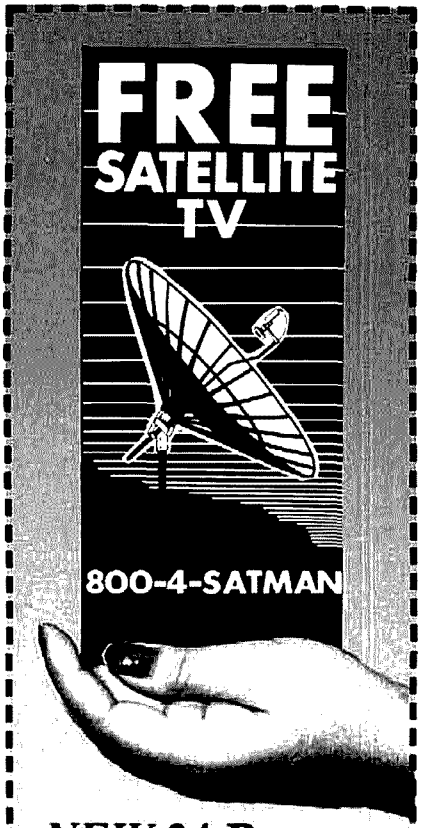
alignment

It should be obvious that the only differences between, say, a 1.0 dB Chebychev and a Butterworth of the same order are small component value differences. Knowing the value of the components is the most important step in getting the filter working.

There are many ways for measuring component

fig. 7. Narrow-band filter CAD program listing.

```
1 REM LC BANDPASS FILTER DESIGN FOR      NARROW BAND FILTERS
3 REM PROGRAMMED BY BOB LOMBARDI WB4EHS
100 PRINT CHR$(147)
105 Q=0:DIM K(7),C(8)
110 PRINT:PRINT:PRINT
120 PRINT "BANDPASS FILTER DESIGN PROGRAM":PRINT:PRINT
122 PRINT"THIS PROGRAM COMPUTES VALUES FOR"
123 PRINT "PARALLEL RESONANT, CAPACITIVELY"
124 PRINT"COUPLED BANDPASS FILTERS":PRINT"THAT ARE (RVON)NARROW BAND(RVOF)
   -(Q>10)"
125 PRINT "ENTER THE NUMBER OF THE DESIRED"
126 PRINT "ROUTINE":PRINT:PRINT
130 PRINT"1 FOR BUTTERWORTH":PRINT
131 PRINT"2 FOR 0.01 DB CHEBYCHEV":PRINT
132 PRINT"3 FOR 0.1  DB CHEBYCHEV":PRINT
133 PRINT"4 FOR 0.5  DB CHEBYCHEV":PRINT
134 PRINT"5 FOR 1.0  DB CHEBYCHEV":PRINT
145 INPUT Q:IF Q<0  AND Q>5 THEN GOTO 135
146 ON Q GOTO 1000,1100,1200,1300,1400
1000 REM BUTTERWORTH SECTION
1010 GOSUB 4000
1015 IF N=2 THEN Q=1.414:K(1)=0.707
1020 IF N=3 THEN Q=1.000:K(1)=0.707:K(2)=K(1)
1025 IF N=4 THEN Q=0.765:K(1)=0.841:K(2)=0.541:K(3)=K(1)
1030 IF N=5 THEN Q=0.618:K(1)=1.000:K(2)=0.556:K(3)=K(2):K(4)=K(1)
1035 IF N=6 THEN Q=0.518:K(1)=1.169:K(2)=0.605:K(3)=0.518:K(4)=K(2):K(5)=K(1)
1040 IF N=7 THEN Q=0.445:K(1)=1.342:K(2)=0.667:K(3)=0.527:K(4)=K(3)
1042 IF N=7 THEN K(5)=K(2):K(6)=K(1)
1045 IF N=8 THEN Q=0.390:K(1)=1.519:K(2)=0.736:K(3)=0.554:K(4)=0.510
1047 IF N=8 THEN K(5)=K(3):K(6)=K(2):K(7)=K(1)
1060 GOTO 2500
1100 REM 0.01 DB CHEBYCHEV SECTION
1110 R=.01:GOSUB4500
1115 IF N=2 THEN Q=1.483:K(1)=0.708
1120 IF N=3 THEN Q=1.181:K(1)=0.682:K(2)=K(1)
1125 IF N=4 THEN Q=1.046:K(1)=0.737:K(2)=0.541:K(3)=K(1)
1130 IF N=5 THEN Q=0.977:K(1)=0.780:K(2)=0.540:K(3)=K(2):K(4)=K(1)
1135 IF N=6 THEN Q=0.937:K(1)=0.809:K(2)=0.550:K(3)=0.518:K(4)=K(2):K(5)=K(1)
1140 IF N=7 THEN Q=0.913:K(1)=0.829:K(2)=0.560:K(3)=0.517:K(4)=K(3)
1142 IF N=7 THEN K(5)=K(2):K(6)=K(1)
1145 IF N=8 THEN Q=0.897:K(1)=0.843:K(2)=0.567:K(3)=0.520:K(4)=0.510
1147 IF N=8 THEN K(5)=K(3):K(6)=K(2):K(7)=K(1)
1160 GOTO 2500
1200 REM 0.1 DB CHEBYCHEV SECTION
1210 R=0.1:GOSUB 4500
1215 IF N=2 THEN Q=1.638:K(1)=0.711
1220 IF N=3 THEN Q=1.433:K(1)=0.662:K(2)=K(1)
1225 IF N=4 THEN Q=1.345:K(1)=0.685:K(2)=0.542:K(3)=K(1)
1230 IF N=5 THEN Q=1.301:K(1)=0.703:K(2)=0.536:K(3)=K(2):K(4)=K(1)
1235 IF N=6 THEN Q=1.277:K(1)=0.715:K(2)=0.539:K(3)=0.518:K(4)=K(2):K(5)=K(1)
1240 IF N=7 THEN Q=1.262:K(1)=0.722:K(2)=0.542:K(3)=0.516:K(4)=K(3)
1242 IF N=7 THEN K(5)=K(2):K(6)=K(1)
```

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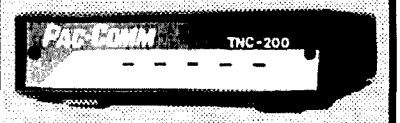
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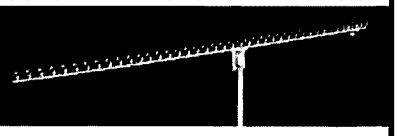
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```
1245 IF N=8 THEN Q=1.251:K(1)=0.728:K(2)=0.545:K(3)=0.516:K(4)=0.510
1247 IF N=8 THEN K(5)=K(3):K(6)=K(2):K(7)=K(1)
1260 GOTO 2500
1300 REM 0.50 DB CHEBYCHEV SECTION
1310 R=0.5:GOSUB 4500
1315 IF N=2 THEN Q=1.950:K(1)=0.723
1320 IF N=3 THEN Q=1.864:K(1)=0.647:K(2)=K(1)
1325 IF N=4 THEN Q=1.826:K(1)=0.648:K(2)=0.545:K(3)=K(1)
1330 IF N=5 THEN Q=1.807:K(1)=0.652:K(2)=0.534:K(3)=K(2):K(4)=K(1)
1335 IF N=6 THEN Q=1.796:K(1)=0.655:K(2)=0.533:K(3)=0.519:K(4)=K(2):K(5)=K(1)
1340 IF N=7 THEN Q=1.790:K(1)=0.657:K(2)=0.533:K(3)=0.516:K(4)=K(3)
1342 IF N=7 THEN K(5)=K(2):K(6)=K(1)
1345 IF N=8 THEN Q=1.785:K(1)=0.658:K(2)=0.533:K(3)=0.515:K(4)=0.511
1347 IF N=8 THEN K(5)=K(3):K(6)=K(2):K(7)=K(1)
1360 GOTO 2500
1400 REM 1.0 DB CHEBYCHEV SECTION
1410 R=1.0:GOSUB4500
1415 IF N=2 THEN Q=2.210:K(1)=0.739
1420 IF N=3 THEN Q=2.210:K(1)=0.645:K(2)=K(1)
1425 IF N=4 THEN Q=2.210:K(1)=0.638:K(2)=0.546:K(3)=K(1)
1430 IF N=5 THEN Q=2.210:K(1)=0.633:K(2)=0.535:K(3)=K(2):K(4)=K(1)
1435 IF N=6 THEN Q=2.250:K(1)=0.631:K(2)=0.531:K(3)=0.510:K(4)=K(2):K(5)=K(1)
1440 IF N=7 THEN Q=2.250:K(1)=0.631:K(2)=0.530:K(3)=0.517:K(4)=K(3)
1442 IF N=7 THEN K(5)=K(2):K(6)=K(1)
1450 IFN=>8THEN PRINT"CAN'T DO 8 POLE 1 DB CHEBUCHEV.  CALCULATING 7 POLE."
1455 IF N=8 THEN N=7:GOTO 1440
1460 GOTO 2500
2500 REM CALCULATION AND DISPLAY ROUTINE
2510 PRINT "WHAT IS THE DESIRED CENTER"
2511 INPUT "FREQUENCY (IN MHZ)";FO
2515 FO=FO*1E6
2520 INPUT "WHAT IS THE 3 DB BW (IN MHZ)";BW:BW=BW*1E6
2530 INPUT "WHAT IS THE INDUCTOR (IN UH)":L:L=L*1E-6
2540 W=2* π *FO
2550 QB=FO/BW
2560 QC=QB*Q
2570 FOR I=1 TO N-1
2575 K(I)=K(I)/QB
2580 NEXT I
2590 CR=1/((W*W)*L)
2600 RE=W*L*QD
2610 FOR I=1 TO N-1
2670 K(I)=K(I)*CR
2680 NEXT I
2690 ON N-1 GOSUB 2700,2720,2740,2760,2780,2800,2820
2695 GOTO 2900
2700 C(1)=CR-K(1):C(2)=C(1):RETURN
2720 C(1)=CR-K(1):C(2)=CR-K(1)-K(2):C(3)=C(1):RETURN
2740 C(1)=CR-K(1):C(2)=CR-K(1)-K(2):C(3)=CR-K(2)-K(3):C(4)=C(1):RETURN
2760 C(1)=CR-K(1):C(2)=CR-K(1)-K(2):C(3)=CR-K(2)-K(3):C(4)=CR-K(3)-K(4)
2770 C(5)=C(1):RETURN
2780 C(1)=CR-K(1):C(2)=CR-K(1)-K(2):C(3)=CR-K(2)-K(3):C(4)=CR-K(3)-K(4)
```

```
2785 C(5)=CR-K(4)-K(5):C(6)=C(1):RETURN
2800 C(1)=CR-K(1):C(2)=CR-K(1)-K(2):C(3)=CR-K(2)-K(3):C(4)=CR-K(3)-K(4)
2805 C(5)=CR-K(4)-K(5):C(6)=CR-K(5)-K(6):C(7)=C(1):RETURN
2820 C(1)=CR-K(1):C(2)=CR-K(1)-K(2):C(3)=CR-K(2)-K(3):C(4)=CR-K(3)-K(4)
2825 C(5)=CR-K(4)-K(5):C(6)=CR-K(5)-K(6):C(7)=CR-K(6)-K(7):C(8)=C(1);RETURN
2900 PRINT:PRINT:PRINT
2910 PRINT "COUPLING CAPS ARE  (IN PF)"
2915 FORI=1 TO N-1
2920 PRINT "C";I;1+1;": ";K(I)*1E12
2930 NEXT I
2935 PRINT
2940 PRINT "RESONATING CAPS ARE (IN PF)"
2945 FOR I=1 TO N
2950 PRINT"C";I;": ";C(I)*1E12
2960 NEXT I
2970 PRINT "THIS FILTER HAS A CHARACTERISTIC Z"
2975 PRINT"OF";RE;"OHMS.":PRINT
2980 INPUT "MATCH TO ANOTHER Z";A$:IFA$<>"Y" AND A$<>"N"THEN GOTO 2980
2985 IF A$="N"GOTO 3100
2990 INPUT"NEW SOURCE AND LOAD Z <50,50>";RS,RL
2992 PRINT
2995 TM=SQR(RE*RS-RS*RS):TN=SQR(RE*RL-RL↑2)
3000 CY=1/(W*TM):CZ=1/(W*TN)
3005 PRINT "ADD";CY*1E12;" PF  TO THE SOURCE END"
3006 PRINT"AND ADD";CZ*1E12;" TO THE LOAD END."
3010 PRINT:PRINT"THEN CHANGE C(1)TO";(C(1)-CY)*1E12;"PF"
3020 PRINT"AND CHANGE C(";N;") TO  "(C(N)-CZ)*1E12;"PF."
3100 PRINT:PRINT
3110 INPUT "ANOTHER FILTER (Y OR N)";A$
3115 IFA$<>"Y"ANDA$<>"N" THEN 3110
3120 IF A$="Y"THEN 120
3200 END
4000 REM FILTER ORDER CALCULATION ROUTINE
4010 REM BUTTERWORTH SECTION
4020 PRINT "ENTER FREQUENCY RATIO F/FC AS A NUMBER, I.E.3":INPUT WR
4025 INPUT"DESIRED ATTENUATION AT F/FC";AD
4030 FOR N=2TO B .
4040 AC=10*(LOG(1+WR↑(2*N)))/(LOG(10))
4050 IF AC=>AD GOTO4070
4060 NEXT:GOTO 4100
4070 PRINT "N=";N;"AT A CALCULATED A=";AC;"DB":RETURN
4100 PRINT"THE REQUIRED ATTENUATION IS OUT OF RANGE FOR A BUTTERWORTH FILTER"
4110 PRINT"OF UP TO 8 POLES."
4120 PRINT"A LOW RIPPLE CHEBYCHEV MAY WORK.":GOTO 130
4130 STOP
4500 REM CHEBYCHEV CALCULATION ROUTINE
4510 DEF FNCS(X)=0.5*(EXP(X)+EXP(-X))
4520 DEF FNASH(X)=LOG(X+SQR(X↑2+1))
4530 PRINT"ENTER THE FREQUENCY RATIO F/FC AS A NUMBER I.E.,3":INPUT WR
4540 INPUT "DESIRED ATTENUATION AT F/FC";AD
4560 EP=SQR(10↑(R/10)-1)
4565 N=2
```


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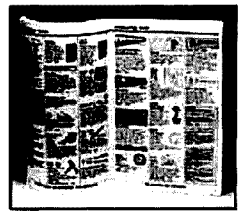
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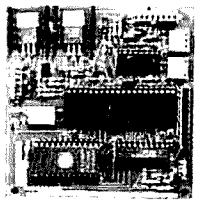
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```
4570 B=(FN ASH(1/EP))/N
4580 WC=(FN CS(B))$WR
4590 CN=(2$WC↑2-1)
4600 AC=10$LOG(1+EP↑2$CN↑2)/LOG(10)
4610 IF AC=>AD THEN PRINT"N=";N;"AT A CALCULATED A=";AC:"DB":RETURN
4620 N=3
4630 B=(FN ASH(1/EP))/N
4640 WC=(FN CS(B))$WR
4650 CN=(4$WC↑3-3$WC)
4660 AC=10$LOG(1+EP↑2$CN↑2)/LOG(10)
4670 IF AC=>AD THEN PRINT "N=";N;"AT A CALCULATED A=";AC;"DB":RETURN
4700 N=4
4710 B=(FN ASH(1/EP))/N
4720 WC=(FN CS(B))$WR
4730 CN=(8$WC↑4-8$WC↑2+1)
4740 AC=10$LOG(1+EP↑2$CN↑2)/LOG(10)
4750 IF AC=>AD THEN PRINT "N=";N;"AT A CALCULATED A=";AC;"DB":RETURN
4800 N=5
4810 B=(FN ASH(1/EP))/N
4820 WC=(FN CS(B))$WR
4830 CN=(16$WC↑5-20$WC↑3+5$WC)
4840 AC=10$LOG(1+EP↑2$CN↑2)/LOG(10)
4850 IF AC=>AD THEN PRINT "N=";N;"AT A CALCULATED A=";AC;"DB":RETURN
4900 N=6
4910 B=(FN ASH(1/EP))/N
4920 WC=(FN CS(B))$WR
4930 CN=(32$WC↑6-48$WC↑4+18$WC↑2-1)
4940 AC=10$LOG(1+EP↑2$CN↑2)/LOG(10)
4950 IF AC=>AD THEN PRINT "N=";N;"AT A CALCULATED A=";AC;"DB":RETURN
5000 N=7
5010 B=(FN ASH(1/EP))/N
5020 WC=(FN CS(B))$WR
5030 CN=(64$WC↑7-112$WC↑5+56$WC↑3-7$WC)
5040 AC=10$LOG(1+EP↑2$CN↑2)/LOG(10)
5050 IF AC=>AD THEN PRINT "N=";N;"AT A CALCULATED A=";AC;"DB":RETURN
5100 N=8
5110 B=(FN ASH(1/EP))/N
5120 WC=(FN CS(B))$WR
5130 CN=(128$WC↑8-256$WC↑6+160$WC↑4-32$WC↑2+1)
5140 AC=10$LOG(1+EP↑2$CN↑2)/LOG(10)
5150 IF AC=>AD THEN PRINT "N=";N;"AT A CALCULATED A=";AC;"DB":RETURN
5200 PRINT"THE DESIRED ATTENUATION IS NOT WITHIN   RANGE OF THIS PROGRAM"
5210 PRINT "FOR THE CHOSEN FILTER TYPE."
5220 PRINT"A HIGHER RIPPLE DESIGN MAY MEET YOUR RE-QUIREMENTS."
5230 GOTO 130

READY.
```

values; a bridge can give quite accurate results, and commercial or homebrew capacitance meters abound. Lacking one of the above, a grid dip meter will do nicely. All you need are some capacitors and coils that will serve as your standards (known values). Using these, an unknown coil or cap can be measured by establishing it in a resonant circuit and measuring the frequency. The *ARRL Handbook*, at least up through the 1982 edition, included a chart for determining LC values using standards of 100 pF and 5μH, although you can use any standard you choose.[4] A frequency on the order of 10 MHz is fine for checking capacitors.

You'll find that the values the program prints out for capacitors are generally not standard and will have to be made up by paralleling one or two with a trimmer. Use the grid dipper to set the value. Likewise, coil values will not work out to an exact number of turns (e.g., 9.39 turns). What you'll have to do, if you're not using a variable coil, is wind the nearest whole number of turns and adjust the coil manually. This is done by setting the grid dip meter to the proper frequency and either squeezing together or spreading apart turns until the value is correct. Fix the turns in place with coil dope. Keep your hands clear while measuring!

Now that the parts are all the correct value, assemble the filter's nodes by shorting all of the caps around an inductor to ground (see **fig. 5**). All of the resulting sections are then tuned to the geometric center frequency, f_o, of the filter. Tune only one adjustment per node! If the parts are properly measured, this should be a very small "tweak"; none may be required below 50 MHz. Connect the filter into its final configuration.

Once the components and sections are tested and assembled, it's best to verify that the filter you have is the one you really wanted (see test setup in **fig. 6**). The signal source can be a cheap generator, either purchased or home-brewed, that puts out several milliwatts of RF. If the RF voltmeter can't read frequency, use a counter. A sweep generator and receiver make an excellent combination, although a sweeper makes some means of accurately measuring the output frequency critical.

If two voltmeters are available, use one to keep the output from the source at the same level. If only one is available, set the source to some level and then sweep it across the filter's passband (still without the filter in line) while recording its level variation for a baseline. Then put the filter in line and sweep the generator across its range, say from the expected 3 dB points, or slightly beyond, while recording levels through the filter. A power meter permits immediate comparison of dB levels; a voltmeter requires a few minutes with a calculator. Once the readings are in dB, plot them versus frequency and see if you get what you expected. Without exceptional shielding in the generator and a more sophisticated detector, it won't be possible to determine stopband response. It's generally valid to assume that if your passband is within your expected limits, the stopband will be also.

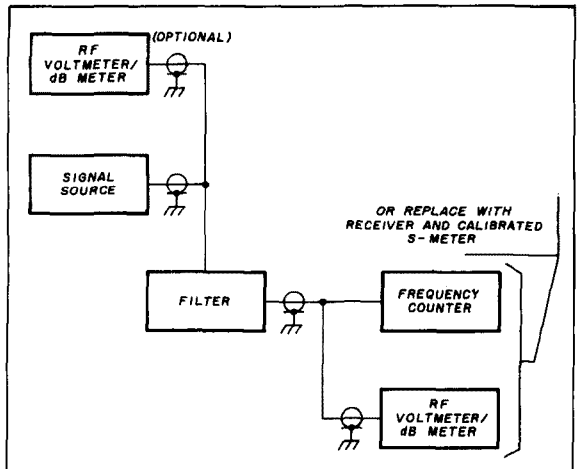


fig. 6. Alignment verification set-up.

If the transmission characteristics of your filter are not as desired, it may be necessary to try to tweak it in. No amount of advice can help here, however, so make sure you've performed all the preliminary steps properly before you start ripping it apart. Many of these filters — especially those built below 30 MHz — work perfectly the first time, . Keep track of what you've done, and recheck it often. If worse comes to worst, retune all nodes and start over again. (If it's any consolation, it does get easier with practice.)

If possible, build some practice filters in the HF range to familiarize yourself with the program and alignment methods. Once you've done that, you should have no problems.

Since the program (**fig. 7**) might be a little long for typing in, I'll be happy to copy it for you. Just send a formatted disk with a self-addressed disk mailer and a check or money order for $3 to me at the address listed at the beginning of the article, and I'll return two verified copies ASAP. Please feel free to contact me about the program or alignment methods. Unfortunately, I can only reply to queries accompanied by SASE.

references

1. Chris Bowick, *RF Circuit Design*, Howard W. Sams & Co., 1982, Chapters 2 and 3.
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ham radio

monolithic RF amplifiers

There's a new building block for HF through 2300 MHz

The monolithic developments that have revolutionized digital electronics at low frequencies are just now being applied to the creation of new types of RF devices. In spite of their novelty, monolithic microwave integrated circuits (MMICs) may soon prove as revolutionary in high frequency techniques as digital ICs have in low-frequency work. At present, entire RF amplifiers are available as MMICs, but soon other RF monolithic devices, subsystems and even entire RF front ends will be built. Silicon and gallium arsenide digital circuitry will also move into the microwave region, making digital RF "front ends" possible. All of these technologies may make the advanced Amateur equipment of the 1990s vastly more capable than today's.

However, even the first RF amplifier MMICs, which are now available, open exciting new possibilities for high performance and small size in Amateur equipment. They are useful in all Amateur bands from HF up to at least 2300 MHz, and should easily find application in receiver and low-power transmitter stages. They are very simple to use; only the DC blocking capacitors and a bias resistor need to be added externally to make amplifiers with bandwidths that can span multiple decades. Furthermore, these MMIC amplifiers now cost less than $3.00 each in small quantities.

This article describes the design of the simplest form of an amplifier stage using one of these MMICs. The amplifier as described is useful as a building block in many types of equipment, and the description provides enough background so that the reader should be able to go on to build more complex, multistage cascades.

MMIC amplifier device characteristics

One company that produces commercially available silicon MMIC parts is Avantek, which offers RF amplifiers in low-cost plastic packages. The main RF characteristics of these amplifier MMICs are summarized in **Table 1**. Avantek offers four versions of their MMIC amplifiers: the MSA 0104, MSA 0204, MSA 0304, and MSA 0404. (The 04 suffix indicates the plastic package style). The four amplifiers provide different gains and bandwidths and four different output power capabilities.

The MSA 01 has the highest gain, but the gain begins to decrease at a lower frequency than for the other three amplifiers. The MSA 01 also has the lowest DC bias requirements, so it's of interest when low current drain is important. Because of its low bias, however, it has a low output power, too.

The next two devices, the MSA 02 and 03, have approximately equal gains. The chief difference between them is that the MSA 03 is specified for higher bias and will provide a bit more output power than the 02.

The fourth member of the family, the MSA 04, has the lowest gain, but the gain is flat to higher frequencies. It also requires the highest bias current and produces the highest output power of the group. **Figure 1** summarizes the gain-versus-frequency responses for the four, each operating at its suggested bias conditions.

I selected the MSA 03 for general-purpose use because it provides reasonably high gain and fair output power capability, and because the lowest current

By Jerry Hinshaw, N6JH, 4558 Margery Drive, Fremont, California 94538

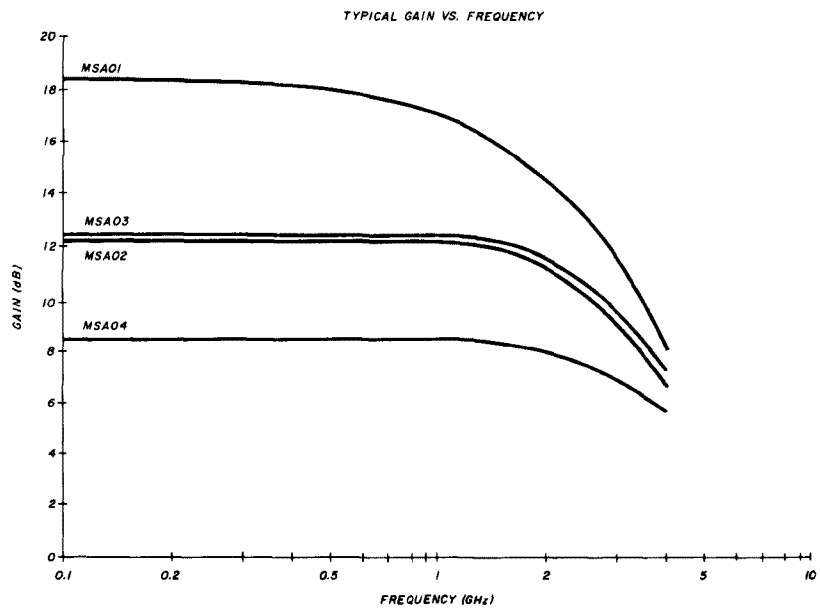


fig. 1. Gain versus Frequency for the four types of MSA amplifier MMICs. Each device is biased at its recommended current level, and appropriate input coupling capacitors are used.

table 1. Summary of MSA 01 — 04 characteristics. The two-digit suffix on each part number indicates package style; the –04 indicates the low-cost molded plastic package.

MMIC MODEL	MSA 01	MSA 02	MSA 03	MSA 04
Gain in dB at 1 GHz	17	13	12	8
Noise Figure at 500 MHz, dB	5	6	6	6
Output Power P1dB at 1 GHz, dBm	1.5	4	10	12

drain is not of much interest in fixed station applications. Therefore, the remainder of this article describes an amplifier built around this device. But the techniques used apply to all four types of MMIC amplifiers, and only the biasing needs to be altered to use another member of the device in the same circuit layout.

MSA 03 monolithic amplifier design

The design of an amplifier block using these monolithic devices is simplified greatly by the fact that no matching needs to be calculated. Maximum gain occurs over a very broad bandwidth with 50-ohm input and output terminations because the device is internally matched. Thus, the only design precautions we must take are to ensure that the device "sees" the proper input and output terminations, namely 50 ohms, and that the correct bias is present.

The first part of this circuit design is satisfied by connecting 50-ohm input and output lines to the package. These input and output connections can be of any desired type — coax, microstrip or stripline, or even waveguide — but the package's leads must be properly transitioned to the connections. The leads are small ribbons, identical to those on transistor packages, and are like those on the commonly-used MRF 901, so they lend themselves to surface mounting on a microstrip circuit board. Microstrip is probably the preferred method of connection, since it's the simplest approach that can give good, repeatable RF performance. So we now have the input and output circuitry "designed." The circuitry consists simply of a random length of microstrip with a 50-ohm characteristic impedance. In microstrip, the impedance of a trace depends mainly on the width of the line, but also upon the thickness and dielectric constant of the substrate material. For commonly available fiberglass epoxy boards, the dielectric constant is about 4.3 in the low GHz range, and so, for a 1/16 inch (1.6 mm) thick board, a trace approximately 0.09 inches (2.3 mm) wide has a characteristic impedance of about 50 ohms. This is a convenient width for construction of the amplifier module because it permits easy connection to output conectors and is a fair match to the widths of many available chip capacitors.

The next step of the simple design process is to

determine the values for the two capacitors needed to provide DC isolation. Although the monolithic amplifier contains most of the circuitry necessary for a complete gain block inside its package, it does not contain the input and output blocking capacitors. These components are not included because it is difficult to make good capacitors of reasonable size right on the silicon integrated circuit chip using monolithic techniques. Therefore, the manufacturer has omitted the DC blocking capacitors, and the designer must provide them externally.

This requirement can, however, be a plus rather than a liability because it allows us to set the low frequency response of the amplifier as we choose. The gain of an amplifier built with one of these monolithic circuits can extend to an arbitrarily low frequency, although not to DC, by the proper selection of capacitor. In other words, the series capacitors used for input and output coupling cause the gain of the unit to fall off fairly rapidly below a certain frequency. This high-pass frequency response is useful in some systems. For instance, it might be that we do not want low frequency gain in the front-end stages of a UHF receiver because excess gain could lead to overloading and distortion caused by signals from strong local broadcast stations.

As a first approximation to calculate the corner frequency of the amplifier, which is what the frequency where the gain has dropped to half its midband value is called, pick a coupling capacitor with a reactance of about 50 ohms at the desired corner frequency. The reactance of a capacitor is equal to

$$X_c = \frac{1}{6.28 \times frequency \times capacitance} \quad (1)$$

Since we want to know the capacitance at a certain frequency where the reactance is equal to 50 ohms, we can rearrange this formula to

$$C = \frac{1}{6.28 \times frequency \times 50} \quad (2)$$

For example, to get a corner frequency of 100 MHz, the capacitors would be approximately 32 pF each, and the gain would be down from the midband value by 6 dB (each of the two coupling capacitors produces a 3 dB drop in the gain of the unit). Note, however, that this formula gives only an approximate value.

Next, the bias decoupling circuitry must be provided. The package used for these monolithic gain circuits has only four leads, and in order to make it possible to connect the grounds with low impedance, two of the leads are internally dedicated as grounds. This means that the two remaining leads are for the input and output, leaving none for a separate bias lead. The bias current for the amplifier must be supplied via its output lead. Thus, it is up to the designer to provide a biasing circuit on the output lead, which is able to supply the neccesary currents to the active stages of the device. At the same time the bias decoupling must not seriously degrade the RF connection of the output — if it does, the gain will be reduced excessively.

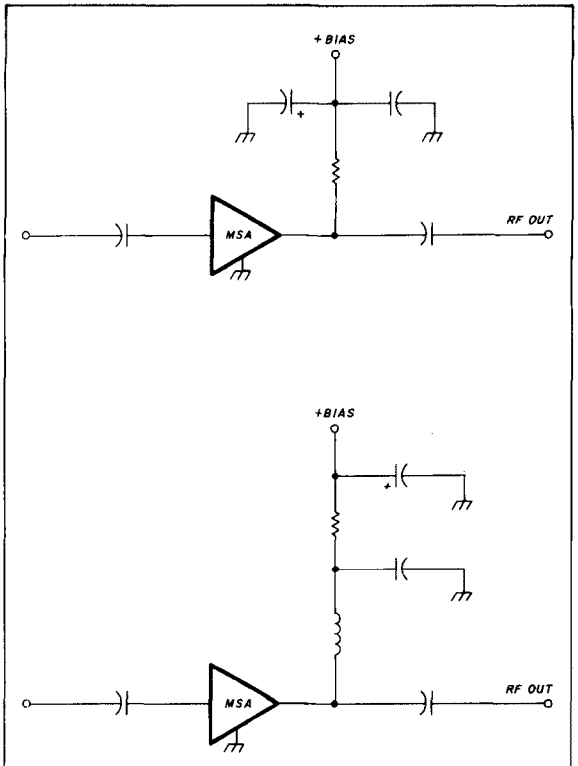


fig. 2. Two bias schemes for MMIC amplifiers. A single resistor can provide adequate performance, but the lower the resistance, the greater the reduction in RF gain. The choke plus resistor can provide maximum gain.

To bias the amplifier we need a circuit that presents a high impedance load at the operating frequencies of the amplifier stage to prevent diversion of RF power from the output into the biasing network. At the same time, however, it must also have low resistance to the flow of direct current. A small inductor fills this need well at high frequencies. The bias circuit must also reduce the supply voltage to the desired collector potential, so a series resistor is needed. **Table 2** gives a few values of resistance that adapt the monolithic amplifier to various supply voltages. If the bias voltage is relatively high, a resistor alone — without the RF choke — will also work well, and will reduce the RF gain only slightly. This is especially true when the resistance needed turns out to be greater than several hundred ohms. These two biasing schemes are shown in **fig. 2.**

table 2. Values of bias resistors for the MSA 03 device for various supply voltages. The power ratings for the resistors include a 2:1 derating factor.

Model	Suggested bias current, milliamps	Supply voltage +12	+15	+24
MSA 01	17	390	560	910
		¼ W	¼ W	½ W
MSA 02	25	270	390	750
		¼ W	½ W	1 W
MSA 03	35	200	300	510
		½ W	½ W	1 W
MSA 04	50	130	200	360
		½ W	1 W	1 W

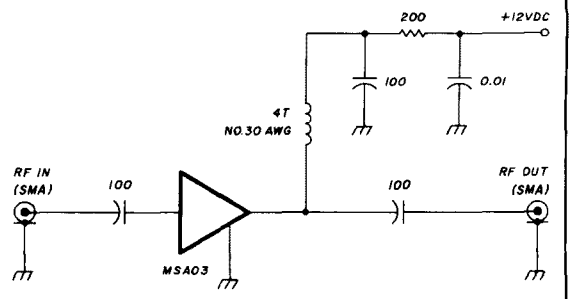


fig. 3. The schematic diagram of the test amplifier described in the text. The 100 pF capacitors are chips, the 0.01 μF capacitor is a ceramic disc. The coil consists of 5 to 10 turns of No. 30 AWG insulated wire-wrap wire wound on a No. 60 drill.

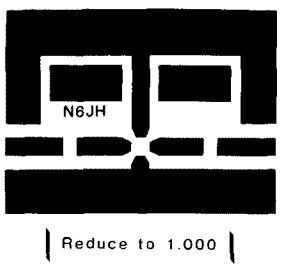


fig. 4. Full size printed circuit negative for the amplifier circuit board. The reverse side of the board is unbroken copper.

Regardless of whether a resistor or an inductor is used to provide bias directly, decouple its far end to ground with a good bypass capacitor to prevent interaction of the external power supply with the RF circuitry. This capacitor, like the RF coupling capacitors in the input and ouput leads, must be chosen with an eye on the operating frequency range. Because we wish this amplifier to work at VHF and above, the capacitors must be suitable and must not have self-resonances below the operating frequency range. If good performance much above 1 GHz is desired, leadless chip capacitors should be used; the more familiar leaded types of capacitors are self-resonant at too low a frequency to be useful here. For VHF and below, though, small ceramic capacitors with short leads work well. With the selection of the coupling and bias circuit elements, the design is done.

construction

This section describes the construction of a single-stage RF amplifier as an example of how useful these gain blocks can be, and suggests what can be expected in terms of actual performance. The amplifier described here is useful as a building block for larger systems or as a test amplifier for use when more gain is needed in a breadboard.

This amplifier was designed to provide useful gain from about 100 to greater than 1300 MHz, using the Avantek MSA-0355-21 device. This device is packaged in a ceramic case, but it is electrically similar to the plastic-package MSA-0304 type. I used it because when I began this project, the plastic devices had not yet been introduced. The device is biased at 35 milliamperes and uses 13-volt direct current. The bias circuit is a simple hand-wound coil, decoupled at the DC end with two capacitors, and a series resistor to drop the 13-volt supply to the required level at the MSA device's output lead. **Figure 3** is a schematic diagram of the test amplifier.

The amplifier was built on a piece of 1/16 inch (1.6 mm) thick fiberglass epoxy, double-clad printed circuit board stock. One side of the board was left untouched so that its uninterrupted copper conductor can serve as the groundplane for the microstrip traces, which are etched or hand-cut onto the other side. **Figure 4** is a full-size negative of the same design that can be used to make an etched circuit board to duplicate the amplifier. The prototype was built on a hand-cut board which has approximately the same pattern of traces as on the negative . Reference 3 describes one method of making hand-cut RF boards.

The components are all mounted to the top side of the board as indicated in **fig. 5**. The input and output coupling capacitors are chips, and are carefully soldered across the gaps in the 50 ohm microstripline leading to and from the device. At the output, a small RF choke connects to a pad on the circuit board which is bypassed to ground at high frequency by a chip capacitor and by a ceramic disc capacitor for lower frequencies. This pad also is the terminus for the dropping resistor, whose resistance depends on the sup-

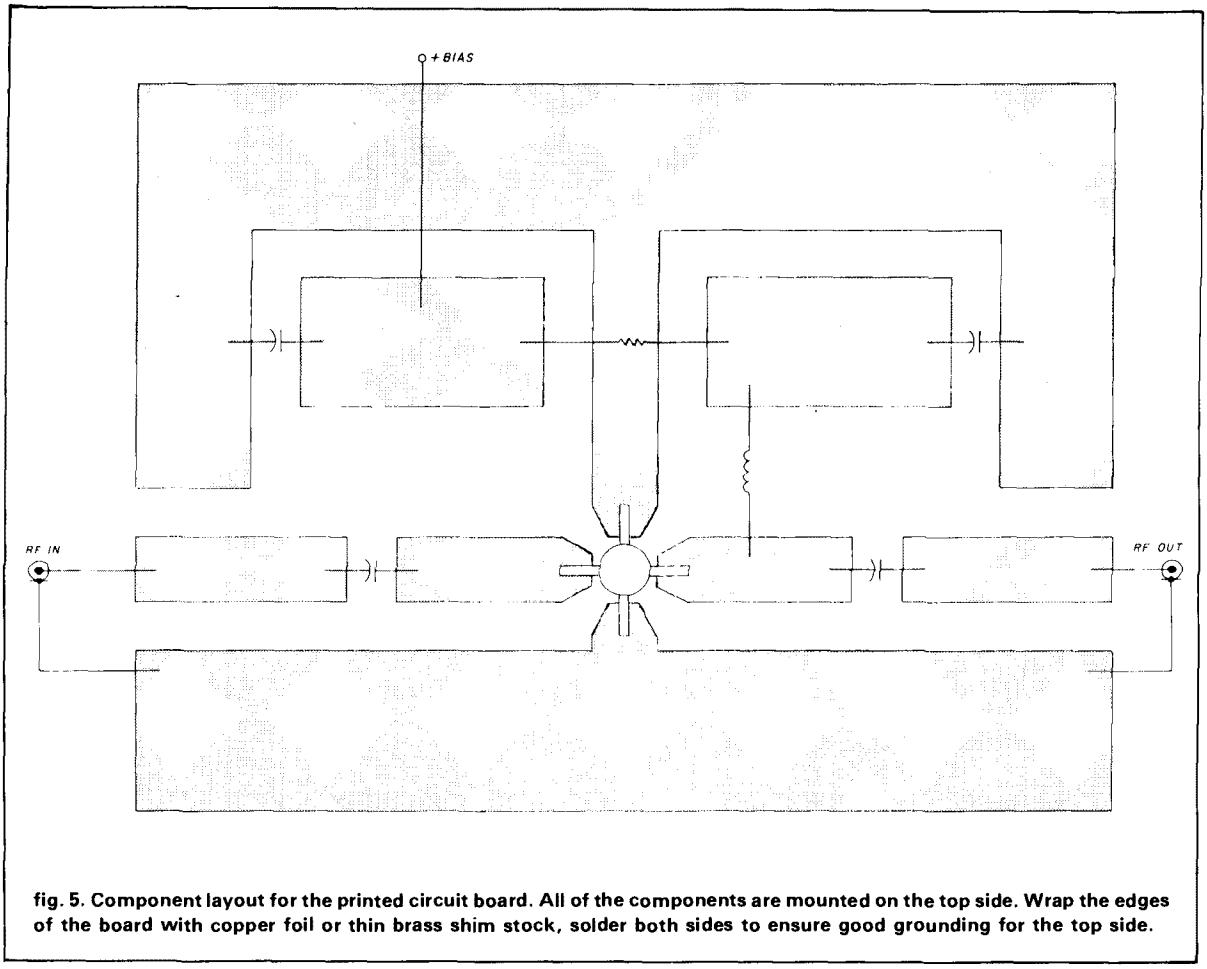


fig. 5. Component layout for the printed circuit board. All of the components are mounted on the top side. Wrap the edges of the board with copper foil or thin brass shim stock, solder both sides to ensure good grounding for the top side.

ply voltage you intend to use. For 13- to 15-volt supplies, 200 ohms is appropriate, and sets the bias to about 35 milliamperes.

The device itself is mounted in a 0.150 inch (3.8 mm) diameter hole drilled through the substrate. This hole provides clearance for the package so that the four leads are flush with the microstrip transmission lines to which they connect. The input and output are soldered directly to their respective lines. The two ground leads are treated somewhat differently; they must be connected to the microstrip groundplane by physically short junctions to reduce their impedance to the minimum. Two methods have been used with good results: the first is to connect the circuit board's top and bottom grounds together with small strips of copper or brass run through the hole drilled to clear the amplifier's package. The ground leads can then be soldered to the top plane; the strips provide a short path to the bottom groundplane. A second approach is to bend the device's two ground leads downward, starting right at the package, and solder the leads to the bottom groundplane. Either method works, but I prefer the former for best high-frequency use and reserve the second technique for lower frequencies, say below 1000 MHz. Whichever method you use, keep in mind that small physical length and good soldering are critical.

Connectors at the input and output complete the assembly. I used SMA connectors so that I could evaluate the performance of the amplifier board at its maximum frequency, but other connectors can be used with good results. Solder the center pins of the connector to the microstripline and the outer conductor to ground, preferably on both sides of the board, but certainly on the bottom side. The clearances can be tight, so be sure that the outer conductor of the connectors does not contact the microstrip line. **Figure 6** is a photograph of the completed prototype amplifier used for evaluating the MSA 03 device,and for the testing procedure described below.

The entire amplifier board has only four capacitors, two connectors, one coil, one resistor and a monolithic amplifier, all mounted on a simple circuit board. *That's all.* Let's see how well it performs.

performance

The data sheet for the MSA 03 indicates that we should expect about 12 dB gain at low frequency and that this gain should drop to about 10 dB somewhere between 1 and 2 GHz. The gain should continue to roll off to about 8 dB at 2300 MHz and down to 6 dB at 3300 MHz. The performance of the test amplifier was measured between 100 MHz and 3 GHz, and the measured data corresponded well to these predicted values. **Figure 7**, as an example, shows the gain between 500 and 1500 MHz, as seen on the screen of a spectrum analyzer. The lower trace is the reference sweep from just the signal generator, while the upper response is made with the amplifier in the line between the signal source and the spectrum analyzer. Therefore, the difference between the two traces is the gain of the amplifier. It is clear that the amplifier is nearly "flat" — that is, its gain doesn't vary by much over this frequency range, but remains a constant 12 to 13 dB.

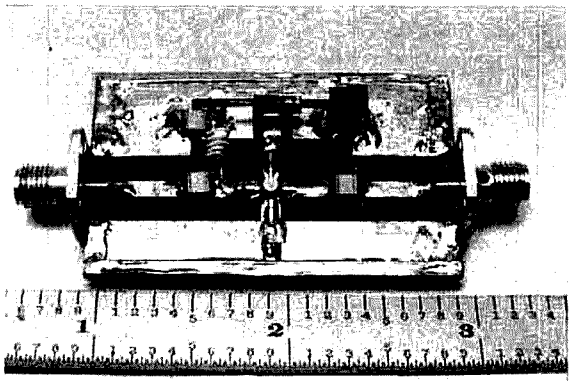


fig. 6. Photograph of the completed amplifier prototype. The ruler scale is in inches. The input connector is on the right.

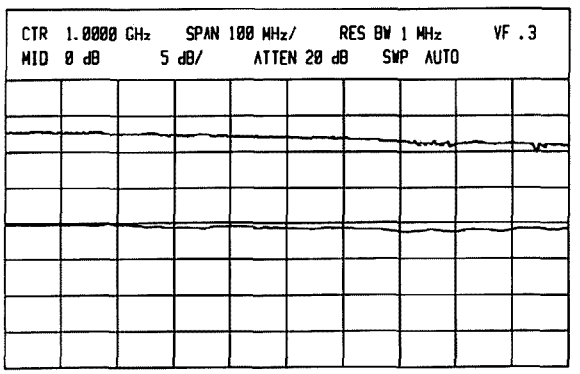


fig. 7. Swept gain response of the test amplifier. The lower trace is the RF signal generator alone, while the upper trace shows the amplifier's gain added to the generator signal. The sweep covers 500 to 1500 MHz, and the vertical scale is 5 dB per division.

Below 500 MHz, the gain remains flat down to the point at which the input and output coupling capacitors begin to significantly reduce the gain, as explained above. **Figure 8** shows the response from 100 to 500 MHz. The gain of the amplifier is down to nearly unity (0 dB) a bit below 100 MHz. The "spike" at the right edge of the trace is a discontinuity caused by the signal generator's stopping its sweep, and is not a resonance in the circuit as it might at first appear to be.

Gain is not the only important amplifier parameter of interest. A second, major item is the amplifier's ability to handle relatively high signal levels. This ability is clearly important in a transmitter chain, and it's also vital in a receiver because the power-handling capability of an amplifier determines the upper end of this dynamic range.

Two common measures of the power handling capability of an amplifier are used. One is the *1 dB compression point*. This is the output power level at which the amplifier's gain is reduced 1 dB from the gain imparted to small signals. This point marks the onset of serious distortion in a "linear" amplifier, and

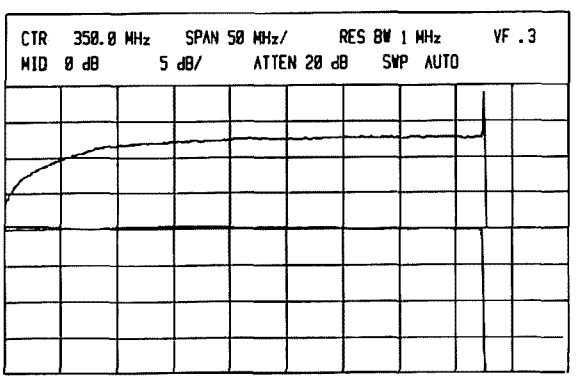


fig. 8. Low frequency gain response of the test amplifier, from 100 to 500 MHz. The gain of the amplifier rolls off as expected at low frequency.

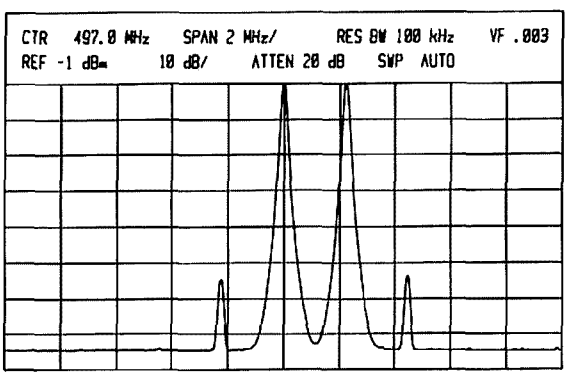


fig. 9. Two-tone intermodulation output of the amplifier.

in general, it is also not many dB from the saturation point of the amplifier. The data sheet for the MSA 03 lists the 1 dB compression point (at 35 milliamperes bias) as about 10 milliwatts for signals below 1 GHz, dropping to about 4 milliwatts at 2 GHz.

A second measure of the power handling capability of an amplifier is the *third order intercept point*, which is the imaginary point at which certain distortion products would be equal in power with the fundamental output of the amplifier. Reference 4 describes all of this succinctly. Practically speaking, third order intermodulation is measured by sending two signals of equal amplitude into an amplifier and observing the output on a spectral display. The display will show the two fundamental signals and will also show (at some level below the fundamental) a pair of responses spaced above and below the two fundmental tones by exactly the spacing between the fundamental signals. These are the third order intermodulation products.

Figure 9 shows this situation clearly. In it, two test tones at about 500 MHz, but spaced by about 2 MHz, have been passed through the test amplifier and the output displayed on a spectrum analyzer. The two fundamental signals are at a level of about −1 dBm each, and the third order products are seen to be about 55 dB below that level. Because the third order products are predicted to rise at 2 dB for each 1 dB increase in the amplitude of the fundamental signals, the four traces would (theoretically) be equal at about +26 dBm. This figure is somewhat higher than expected, as the data sheet suggests a third order intercept point of +23 dBm for this frequency and bias condition. However, the measurement is subject to some error, especially in the precise measurement of power level, so this test indicates that the amplifier's performance is approximately equal to the data sheet's predictions.

conclusion

These MMIC amplifiers offer a powerful new flexibility to RF designers; their cost is low enough so that Amateur experimenters can afford to use them. They make it possible to build high frequency equipment more easily than ever before, and with better results. I predict that you'll be seeing many of them in designs over the next few years, and that they will open up the microwave bands to a lot of new operation.

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vswr bridges

Simple device requires careful design for accurate, reproducible results

Some conventional VSWR bridges have a bad reputation; there's much confusion. They may have a fairly flat forward output vs. frequency response and even a good reflected null at a given frequency, but at high known VSWRs, they don't seem to do the desired job. Sometimes the reflected power is even larger than the foreward! Now — in this era of transistorized amplifiers, some don't seem to provide the necessary protection. It would appear that their design and construction, therefore, need more attention.

what's a bridge?

Conventional VSWR bridges are devices that sample voltage and current components to a load. They all use a voltage divider and a current divider, although sometimes these are combined to look more like a pickup loop or short transmission line. The sum of these normalized samples is the forward output voltage. Usually RF detectors are built-in to provide DC outputs. It sounds simple, and it is; but it works as accurately as possible to produce the samples, and then independently adds or subtracts them. These components enter into the basic definition of forward or reflected power.

why so much discussion?

Obviously, there are several pitfalls — in design, construction, and measurement interpretation. If there weren't, there wouldn't be so many articles on the subject or so much discussion about it. Bridge circuits are often evaluated in terms of reflectivity, where the ratios are between the forward and reflected output voltages, usually expressed in dB, and by forward voltage vs. frequency measurements. A troublesome bridge may have an honest reflectivity ratio of only 14 dB, yet the user may think it's better! Let's examine some of the reasons for this.

how good?

A reflectivity ratio of 14 dB corresponds to a VSWR ratio of 1.5:1, so measurements would then have a corresponding uncertainty. A true 5:1 forward-to-reflected voltage ratio is a 25:1 power ratio, or 14 dB reflectivity. Most practical bridge circuits should reliably hold a 26 dB reflectivity ratio, corresponding to a 1.1:1 VSWR uncertainty, in normal production practive. (Often a lowpass harmonic filter is used after the bridge, adding its 1.1:1 VSWR, too. A lab instrument can hold 35 dB, and with further adjustment, 50 dB.

diode knee — a big deal!

The measurements will vary with temperature and/or level because of the diode's offset voltage. This threshold of the diode will suppress a small reflected voltage and make one think that the bridge is much better than it actually is. Even with "hot carrier" silicon diodes, this is a factor. Forward-biasing the diodes helps, but may substitute the problem of having an output even if there is no input signal. Nonlinearity of the scale calibration at low signal levels is usually a "diode knee" effect, and therefore temperature-sensitive. It's best, then, to be at a level high enough that results aren't affected by the diode knee. A diode threshold of 0.5 volts means that any reflected samples below that won't give any output indication.

good ground presents bigger problem

Many construction problems are partly ground-oriented. One must remember that for RF purposes, the concept of a ground exists only as a reference point for one's thinking. RF impendance and transmission line effects limit the usefulness of that concept. Often, it's no more than an aid to sloppy thinking, because of the confusion it creates. The "ground" lead of every component is a separate RF ground, but all the common resistance and inductance — which counts greatly in circuits such as these bridges — doesn't show on the schematic. Where the ratios between signals are very large, the ground location must be correspondingly large.

By Will Herzog, K2LB, 406 Yarmouth Road, Rochester, New York 14610

sampling voltage not easy

The voltage divider's bottom capacitor's lead must be the reference ground. Normally the capacitive reactance of this capacitor is low, to present a source impedance to pump the current transformer up and down, by the voltage sample amount. Some designs use a voltage transformer instead of a capacitative divider; these designs can represent an even more difficult design problem for phase control at higher powers. In most cases, the detector level is well below the RF power level, so all RF power line and ground currents must be kept out of the voltage sample and the detector. The detector's ground is the bottom of the filter capacitor that follows the rectifier diode.

voltage divider compensation

The inductance in series with the voltage divider's bottom capacitor must be as small as is practical. A small amount may be compensated for by placing an inductor in series with the top capacitor so that the resonant frequency of the top capacitor and the added inductor are equal to that of the bootom capacitor and its (minimal) lead inductance. Previously, attempts were made to keep the stray inductance of the top capacitor short as well — yet this teaches that longer can be better!* The filtering for the output leads also must not insert ground currents into either the detectors or into the voltage sample. A 100-microhenry choke is an appropriate value to use for 2-30 MHz because its parallel resonance (i.e., the highest impedance or best filtering point) is about 12 MHz, the center of the band. Often we shield or filter these output leads, but forget about where the bypass capacitors or shield(s) are grounded.

current sample

Most current transformer designs use an electrostatic shield to prevent capacitive coupling from the coaxial line to the current sample, but forget that the capacitance from both ends of the sampling inductor and its load resistors to ground must also be small, so the voltage sample doesn't inject an extra component into the current sample.

current transformer paradox

There is an intrinsic error caused at the low-frequency end by the inductance of the secondary. The primary magnetic field in the transformer is directly in phase with the actual primary current. Any delay in the field's buildup is an inductive effect that would retard the current. This delay is a consideration in a voltage transformer, but in a current transformer the current itself is the thing to be measured or referenced. The impedance of the primary must be kept small enough to not affect the current. So the magnetic field builds up to nearly the speed of light. But it's the rate-of-change of this field that induces the internal secondary voltage. Note that this is the theoretical open circuit voltage. It's not the same as the terminal voltage that we can see, measure, and use. For our sine wave current, the phase of this induced voltage must be at a maximum at the same time that the primary current is going through zero; thus, there's always exactly 90 degrees between them. The induced voltage lags the current by 90 degrees (or more conventionally stated, the current leads the induced voltage by 90 degrees). The actual terminal voltage is then further delayed to nearly a 180-degree total because a second, never-quite-90 (let's assume 88) degree phase shift adds to this to make an overall 178-degree current lag in the dividing transformer. The 88-degree shift is caused by the inductive reactance of the winding, its resistance, and the load resistor.

current sample termination resistors

We then also have a 2-degree leading current sample from the second terminating resistor, since the reference ground was considered as the junction of the two resistors. Note that if the outputs are to be equal, the resistors for each polarity of current samples must be equal. It's nice to have highly stable 1 percent resistors. This 2-degree error is caused by insufficient inductance of the secondary itself. So this inductance must be large compared to the termination resistors so that we get as close as practical to a 90-degree phase shift of the current lagging the induced voltage.The actual error is the angle determined by the arctangent of the ratio of the secondary's reactance, at the lowest frequency, to the terminating resistance.

it can be compensated

One can look at it as though there's a constant current driving the transformer at the time when the sample should be; but there's a small phase lag of the actual current driving the termination. Shunting the termination resistors with a second series combination consisting of a resistor and capacitor can correct for this phenomenon and for core loss! This works by modifying the termination by unloading and presenting a slightly capacitive load at the lowest frequencies.*

current sample terminations

The required turns ratio for the current transformer is determined by considering the practical value of terminating resistors for the output voltage desired, and by how much power is to be wasted in them. To sim-

plify the current transformer design, one might think that lower values of ohms, such as 5 ohms, would lead to better bridges, but the output must then be kept low because of the power allowable in these resistors; but then the desired output voltage is low, and that leads to diode-knee effects and problems of temperature and stability. All designs are a compromise, with time and money determining when to shoot the engineer and ship the radio.

transformer phase delay

Any tansformer has a phase delay of its output compared to its input because of the transmission line effects of its wire. This limits the bandwidth for most designs. I measured the phase delay of several transformers by using a vector voltmeter, with its reference probe across a 10-ohm resistor threaded through the core (lead length must be 1/4 inch maximum) and the other probe across the normal termination resistor. Ten to 30 degrees seems typical at 30 MHz for 2 to 30 MHz designs. But one commerical design had a 180-degree shift at 35 MHz!

two are better than one

One way of reducing the transformer phase delay and helping to alleviate some of the problems described above, is to use two transformers, connected in series, rather than only one. The first core encircles the transmission line, and its output is simply passed through the second core to be its one-turn primary. The second core then requires fewer turns because the effective turns ratio is the product of the turns on each transformer. So two 7-turn transformers connected in series give the same ratio as one 49-turn transformer. The total length of wire required is then less, lowering the total phase delay. The two-transformer design also allows the link-turn to be grounded to help act as a second electrostatic shield to increase the "common-mode rejection."

current sample compensation

I propose a way to correct for the transformer's phase delay by adding an extra inductor in series with the terminating resistor (about 0.02 to 0.1 microhenries in series with 56 ohms at 30 MHz), so the current-sample voltage seen by the detector diode is not only the transformer's secondary current through its termination resistor, but also has an extra leading voltage component because of the added inductor. This works because the voltage across and inductor leads the current through it to shift the current sample's phase just enough to make up for the transformer's delay.* Previous procedures called for reducing the inductance of the current transformer's terminating resistors — I'm saying to increase it, even build it in! Quarter-inch lead length variation can be significant.

tolerance adjustment handle

A variable resistor across the current sample, to shunt some of it, can serve as a stable null adjustment control. Low inductance film types are now available. Most variable capacxitors that have enough range are quite expensive, and are either large in size or subject to drift with temperature.

misinterpreting the results

Here's how I kid myself into making a bad bridge look good for a test. Let's assume that I have the normal setup of a (transmitter) generator driving the voltage sample divider, then into the current transformer, and then to the (antenna) design nominal load, usually 50 ohms. The forward output is the largest output side, where the two equal samples add, and the reflected output is the null output, where the two equal samples subtract. The samples are usually diode detected to provide DC outputs to the forward and reflected meter(s). The usual test for the bridge is to set the null, making the two samples equal at that point by shifting the amplitude of one of them. I assume that the two current samples must be equal also. But because of the diode knee, the fact that this null might not be very deep is usually hidden, and we usually accept some reflected power at this null because there is no phase control. I assume that the two samples are really equal, and I assume that they both hold constant with frequency. If they aren't, I fix it, so I succeed in making the bridge work into 50 ohms (because I've nulled/gimmicked/set it for the desired result at this point). But when I use this bridge, these errors could be great enough for me to be bothered by the forward power variations not holding power, and the reflected power indications could become virtually useless. So one can't extrapolate from this simple alignment-test that the bridge will perform at other frequencies or with various loads.

bridge makes a VSWR

At the high frequency end, the shunt capacitance of the voltage sampler and the series leakage inductance of the current sampler and its leads cause the VSWR into the bridge itself to never be the same as the load presented to the bridge. This is a larger effect when the bridge is a low-power one, but is mostly a function of layout and can usually be compensated for by adding a small series inductance at the input to complete a "T" lowpass filter. For fair comparison

*Patent applied for, Harris Corporation, 1680 University Avenue, Rochester, New York 14610.

or evaluation, directional coupler bridges should not be put in series and read all at once.

is it symmetrical?

The usual way to check the forward sample's null is to run the bridge backwards — i.e., reverse the coax connections and the metered outputs to see that the bridge is symmetrical. Actually, the impedance to the coax ports should be the same both ways, but the reflected null outputs should not! The backwards null should indicate higher because the impedance at that point really isn't the same. The presence of the voltage sample, or the current transformer, makes it different. So this effect will cause even a perfect bridge to be one-sided at its high frequency end, if the null is checked in both directions. This presents a potential interpretation propblem, and explains why input and output labels should be used. This effect is greatest for low-power units.

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modifying the Yaesu FT-301
for 30-meter coverage

Five straightforward steps convert this transceiver for 10-10.5 MHz operation

The Yaesu FT-301 is an attractive looking modern transceiver marketed in the very late 1970s and early 1980s. It was built in four basic models identified as the FT-301, FT-301D, FT-301S, and FT-301SD. The suffix D denotes the model with a built-in frequency counter which drives a digital frequency readout. The suffix S denotes the QRP version, for which the final power amplifier is not included. It uses the driver as the output amplifier, and is rated at 10 watts output. The full power version was designed to run about 230 watts input, and is rated at 100 watts output.

The FT-301's operating capability includes SSB (upper and lower sideband), CW, AM, and FSK. The rig is all solid-state, including the final amplifier, and has provisions for the usual crystal filters for controlling receiver bandwidth. Also included is a noise blanker, an optional RF speech processor, marker generator, RIT, and provision for an external VFO option. The transceiver operates on 13.5 volts, and draws 21 amperes on transmit (3.1 amperes for the S model), and 1.1 amperes on receive.

The transceiver covers the Pre-WARC 10 through 160-meter bands (the 10 meter band is divided into four 500 kHz segments designated 10A, 10B, 10C, and 10D), and one additional position on the bandswitch designated AUX. It is this last position that can be converted to provide coverage of the range 10.0 to 10.5 MHz, including the 30-meter Amateur band from 10.1 to 10.15 MHz.

The AUX position was originally intended by Yaesu to be an option that could provide coverage of the 27-MHz Citizen Band. Altering it to cover the 30-meter band is not difficult, but does require some care and patience. Fortunately, the conversion can be accomplished with the tools and test equipment found in most Amateur stations. Conversion has no effect on the operation of the other bands in the transceiver.

The conversion project is divided into five steps, one of which is not required in the dial (nondigital) models. These steps cover conversion of the digital display PB-1542; the band oscillator crystal unit PB-1441; the bandpass filter unit PB-1442B; the tunable filter trimmer unit PB-1446; and the output low-pass filter unit PB-1445. (The PB designations above are the same as shown in the documentation included in the Yaesu instruction manual for the FT-301.)

overview of the transceiver block diagram

Understanding the principles employed in the design of this transceiver will make the description of the conversion steps easier to follow.

In the transmit mode, the signal originates at 9 MHz, where crystal filters and a balanced mixer provide upper and lower sideband capability, and where provision is also included to generate CW, AM, and FSK signals.

The 9-MHz signal is heterodyned to the desired output operating frequency by being mixed with a pre-mix frequency in a 500-kHz segment located between 13.9985 MHz and 39.0 MHz, with its exact location dependent upon the band in use. This pre-mix frequency is itself the heterodyne product of two frequencies set by the main tuning VFO (between 5.0 and 5.5 MHz), and a band-determining crystal. (An exception is the 80-meter band, for which the 5.0-5.5 MHz VFO output is used directly as the pre-mix frequency.) **Table 1** shows the data as presented in the Yaesu instruction manual, to which a final line has been added showing the proper frequency relationships for 30-meter band coverage.

The same pre-mix frequency is used in the receive mode, and several of the same tuned circuits are switched so that they serve the same frequency-determining function for receive as for transmit. Consequently, one set of modification steps serves for both transmit and receive, and essentially the same

By Warren Offutt, AF9Q, 1012 Hawthorne Lane, Geneva, Illinois 60134

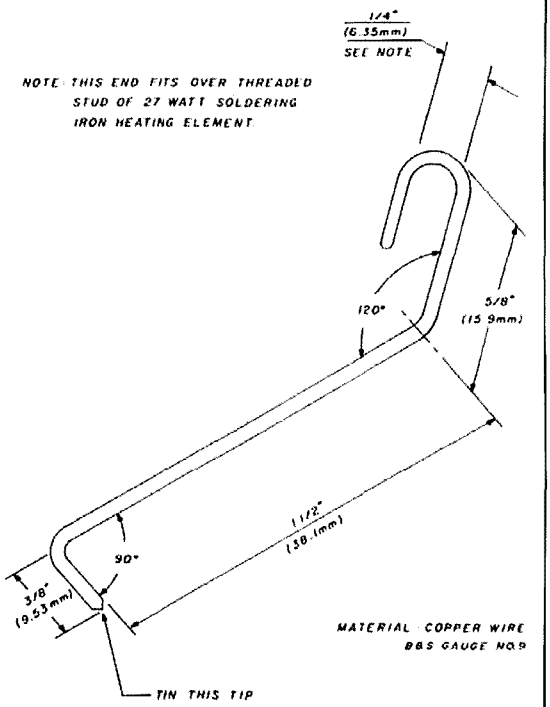


fig. 1. Soldering tool.

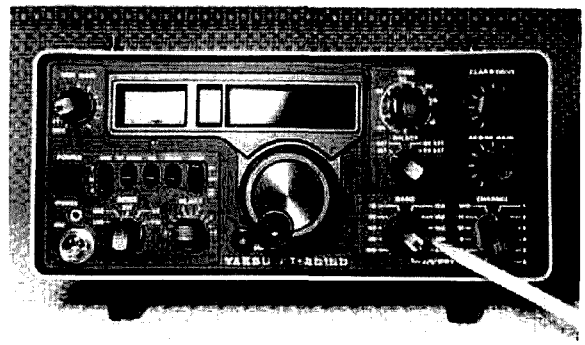
table 1. Frequency relationships.

band	frequency MHz	crystal MHz	premix out MHz
160 meters	1.5- 2.0	16.0	10.5-11.0
80 meters	3.5- 4.0	none	5.5- 5.0
40 meters	7.0- 7.5	21.5	16.0-16.5
20 meters	14.0-14.5	28.5	23.0-23.5
15 meters	21.0-21.5	35.5	30.0-30.5
11 meters	27.0-27.5	41.5	36.0-36.5
10 meters A	28.0-28.5	42.5	37.0-37.5
10 meters B	28.5-29.0	43.0	37.5-38.0
10 meters C	29.0-29.5	43.5	38.0-38.5
10 meters D	29.5-30.5	44.0	38.5-39.0
WWV	5.0	13.9985	13.9985
30 meters	10.0-10.5	24.5	19.0-19.5
	VFO: 5.0-5.5 MHz	IF: 9.0 MHz	

steps that align the receiver also align the transmitter.

The front panel band-peaking control adjusts a set of three permeability-tuned circuits, which resonate the receiver mixer, RF stage, and transmitter driver. Each band has its own set of trimmer capacitors. The permeability-tuned coils permit the user to peak the tuned circuits for the band in use, and also set the proper L/C ratio over the 1.5-30 MHz frequency range.

Finally, the output of the transceiver is connected to the 50-ohm antenna coaxial receptacle through a



AUX setting of bandswitch used for 30 meters.

set of low-pass filters, with the proper filter selected by the bandswitch for the band in use.

conversion steps

The five conversion steps may be tackled in any order you prefer; however, the order described below is recommended. In this order, you'll become familiar with the transceiver's interior by accomplishing the easier tasks first. None of the steps is unreasonably difficult, nor is any likely to cause problems with adjustment. You will have to fabricate one special soldering iron tip, which is described below. The conversion is not for the rank beginner in electronics, but you don't need to be a PhD either. Carefully follow the description and instructions below, work slowly, and you'll be rewarded with very attractive results. (The work should take about five or six hours.)

parts and tools needed

Only ordinary tools are needed for completion of this project. They include the following:

- miniature wire cutters and needle-nose pliers
- electronics grade forceps, with locking handle
- assorted small screwdrivers with conventional flat blades
- plastic hex alignment tool to fit T-1006 in the bandpass filter unit
- small (27 watt) soldering iron, with screw-on tip (Ungar or Radio Shack)
- medium-size soldering iron (approximately 60 watts)
- flashlight and magnifying glass (or magnifying glasses you can wear)
- proper size Phillips screwdrivers (see below)
- grip-dip meter (optional, but desirable)
- dummy load (50 ohm)
- RF wattmeter or equivalent power meter

Be sure the Phillips screwdrivers really fit the screw heads properly. When they are correctly matched, the screwdriver securely grabs the screw; when they are

poorly mated, it is impossible to crack free a tight screw. *Don't bypass this step.* Four of the screws you'll have to remove are hard to reach, and have been glyptal-locked in place. You'll need a good match between the screwdriver and the screw head. It seems to me that Japanese screws have shallower wells than the American screwdrivers can accommodate, and it's sometimes necessary to file or grind off a little of the screwdriver point in order for it to seat properly.

The grid-dip meter is optional, but it is very handy for checking and preliminary alignment.

You'll have to fabricate a special tip for the soldering iron. Figure 1 shows its design. A convenient source of hard copper wire of the proper gauge is the copper rod sold in hardware stores for replacing the pull rod in a ball-type toilet tank valve. It is 105 mils in diameter, and is equivalent to B&S 9 gauge wire.

Materials for the project include the following:

HC-25U crystal, 24.50000 MHz

5 signal diodes, 1N914 or equivalent

1 good quality fixed ceramic capacitor 15 or 20 pF

2 good quality fixed ceramic capacitors 45 pF

3 miniature compression trimmers, 200 pF

miscellaneous hookup wire, insulating tubing, etc.

The crystal should be ordered for a 30-pF load capacitance, fundamental frequency of 24.50000 MHz, parallel resonance, and 0.001 percent accuracy.

Standard width (5/8-inch/1.59-cm) compression trimmers may be used, but if you can find the narrower version (3/8 inch/0.95 cm), such as Calectro Catalog No. R1-248, 55-300 pF), installation is easier. The capacitance when resonated will be between 100 and 175 pF.

preparation

Disconnect all cables from the FT-301 transceiver, and remove the top cover by snapping out the four cover locks. Lift the cover free, but be careful not to tear free the speaker cable. It has an in-line connector that may be released to permit setting the cover aside in a safe place.

Turn the transceiver over, placing it on a soft protective surface. Remove the twelve small screws that hold the bottom cover in place (four on the bottom, four each on the left and right sides), and carefully lift the cover free and place it aside.

Position the transceiver with the front panel toward you (and the rear apron away from you), upside down, so you're looking into the bottom of the chassis. This is the position in which steps 1 through 4 are completed, and is the position to which directions below apply.

Inspect the transceiver to familiarize yourself with the general layout. Make notes to assist in reassembly when you're finished. It goes without saying that you should be satisfied that the rig works properly before you begin; no modification should ever be started unless the equipment is in good working order to start with. Otherwise, if problems arise, you may never know their true cause.

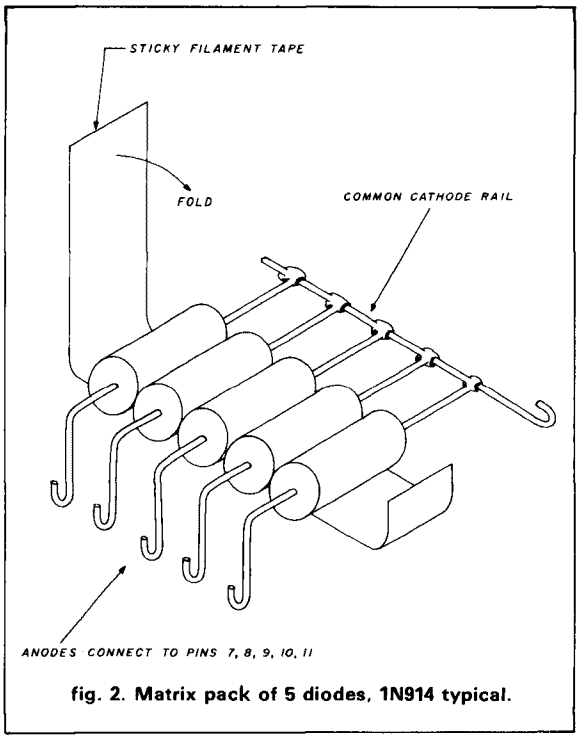


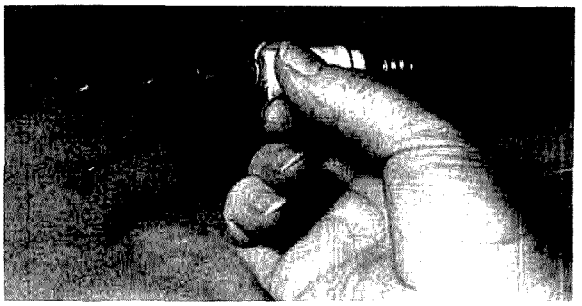
fig. 2. Matrix pack of 5 diodes, 1N914 typical.

step 1: conversion of the digital display

If the model you are converting does not have the digital display, but is instead the analog dial model, skip this step and go directly to step 2.

Locate the display logic unit, PB-1542 — the vertical printed wiring board, immediately behind the center of the front panel, and parallel to it. It is the board on which the LED digit displays are mounted. Along the bottom edge of this board (which is now "up" because the rig is upside down), you'll find a row of 15 terminals, part of a 15-pin receptacle/plug combination which connects the board to a small cable harness (which itself has an in-line receptacle/plug combination 1 inch (2.54 cm) or so further toward the rear). Counting from your left, identify pins 7, 8, 9, 10, and 11. (Pins 14 and 15 have no wires connected to them; be sure you count correctly.) These five pins must be pulled to ground when in the AUX band position so that the display reads 10 to the left of the decimal point.

A few words of explanation may be helpful at this point. The digital display is not entirely a genuine frequency counter. The digits to the left of the decimal



Assembly of soldering tool behind screw-on tip with old tip end removed.

point are "forced," in that a diode matrix activated by the bandswitch commands a display of 28 for 10 meters, 21 for 15 meters, etc. The diode matrix is located in the counter mixer unit (the square box a few inches further toward the rear). Unfortunately, the diode matrix is potted in a 14-DIP package, and altering the 27-MHz section to yield 10 is not possible. We shall use the five diodes to construct an external diode matrix, which will have the same effect as reprogamming the matrix in the DIP package. As a matter of fact, it's easier to do the job as described below than it would be if it were possible to reprogram the DIP package because no disassembly of the counter mixer unit will be required.

Form the five diodes in a small pack as shown in **fig. 2**. The cathodes should be connected together as a common bus. Carefully position the diode pack on the up-facing surface of the 15-pin receptable, and connect the five diode anode leads to pins 7, 8, 9, 10, and 11. (Pin 11 causes the first digit to be 1; pins 7, 8, 9, and 10 cause the second digit to be 0.) With care (watch out for solder bridges!), solder the five diode leads to the correct pins. Inspect your work.

Now, locate the black wire, the far end of which connects to the AUX lug on the bandswitch wafer (SW2A-1 in the Yaesu documentation) closest to the front panel. If your rig doesn't have the 27-MHz option installed, the near end of the black wire will be found connected to pin 13 of the same group to which you just connected the diodes. If your rig has the 27-MHz option installed, the near end will be found connected to one of the center lugs (pin 6, but double check) of the cable in-line receptacle-plug combination mentioned above. When you've located the correct wire, disconnect its near end from whichever place you found it, and connect it now to the common rail of the new five diode matrix you just installed. The purpose is to ground the common bus of the five diodes when the bandswitch is in the AUX position. Using some sticky filament tape or equivalent, fasten the five diodes down neatly to the top of the 15-pin receptacle.

You've now completed the conversion of the digital display. You may check your work by connecting the power supply to your rig and turning it on. With the bandswitch in the AUX position, the display should operate normally, and the first two digits should read 10. Before you consider this step completed, run through the other band positions, and verify that the display reads the correct digits to the left of the decimal point for each band. If not, the likely cause is either that you connected to wrong pins, or (more likely), that you have a short or solder bridge between two pins. You should not encounter such problems if you work very carefully and take your time. It's very satisfying to see that 10-MHz frequency displayed, but don't be fooled: the rig isn't yet on 30 meters! Carefully redress the wires around the display unit to their original positions, and go on to step 2.

step 2: conversion of the band oscillator crystal unit PB-1441

This one is easy. On your left you'll find a flat horizontal metal shield plate. It has access holes for various alignment trimmers and is held in place with eight screws. Remove them, noting their position carefully (they're not all the same length). Remove the flat shield and lay it aside in a safe place.

A few inches behind the front panel, in the first compartment under the shield, you'll find a crystal socket board with eight, nine, or ten crystals in it. Eight crystals are for the existing bands, 160 through 10 meters. The ninth crystal (if present) is for the JJY/WWV option. The tenth crystal (if present) is in the AUX socket and is for the 27-MHz option.

Incidentally, if your rig does not have the option to receive WWV on 5 MHz, this would be an ideal time to add it. The WWV crystal should meet the same specifications mentioned above and have a frequency of 13.99850 MHz. The WWV trimmer closest to the front panel tunes the circuit for oscillation; the WWV trimmer one step toward the rear apron adjusts for zero beat with WWV. This option works well only with the mode switch in the LSB position.

The crystal socket in the rear row, closest to the center of the rig, is the proper place for the new 24.50000 MHz crystal for the 30-meter band. You may install it now.

It will be necessary to pad the existing AUX trimmer with a 15 to 20-pF capacitor. The easiest way is to form the leads of the capacitor so it can lie alongside the AUX trimmer, and carefully tack-solder the leads to the AUX trimmer leads. The proper two AUX trimmer leads are (a) the lead facing the rear of the chassis, and (b) the lead facing the center of the chassis. Work with a clean, hot iron, and pre-tin the new capacitor leads. Don't overheat the trimmer. This task is easy if done with care.

Some of the subsequent steps depend on this circuit oscillating properly. The likelihood is that correct oscillation will be obtained if the AUX trimmer is set at mid value. However, a simple additional test can be used to verify proper operation as follows.

The sixth harmonic of 24.5 MHz is at 147 MHz, which falls conveniently in the 2-meter band. Assuming you have (or can borrow) a 2-meter HT, tune it to 147 MHz, and then listen for the squelch to open when you tune the AUX trimmer. (Power must be applied to the FT-301, of course, and the bandswitch must be in the AUX position. The FT-301 should be in the receive mode.) With the HT antenna within a few inches of the crystal compartment, the 147-MHz signal should be at full quieting. The proper setting of the AUX trimmer should be about 1/4 turn less capacitance than that at which oscillation starts; in other words, don't crowd the maximum capacitance setting at which the circuit begins to oscillate. You may, if you wish, use the HT to judge the position at which oscillation is strongest.

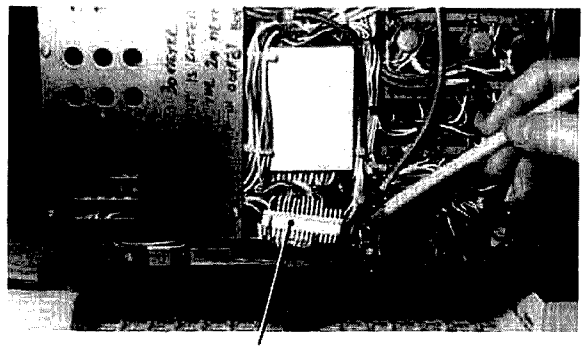
When you're satisfied with the oscillator operation, you may proceed to **step 3**. (Note: you may find that the 24.5-MHz crystal oscillates just fine without the extra 15 or 20-pF capacitor. If so, the extra capacitor may be omitted; however, to be sure you are optimally tuned, it's a good idea to add it, especially since it's easy to do.)

step 3: conversion of the bandpass filter unit

The compartment next to the crystal compartment of **step 2** is the bandpass filter unit PB-1442B. It is located one compartment away from the crystal compartment, toward the rear of the transceiver. Within this compartment, locate bandpass filter T-1006, which is toward your left and toward the rear. It is the filter that connects to the AUX lugs on bandswitch wafers SW2A-6 and 7. The primary connection has a piece of hookup wire from the AUX lug on wafer 6 (an easily accessible lug toward the center). The secondary connection has a piece of shielded cable that connects to the AUX lug on wafer 7 (also easily accessible).

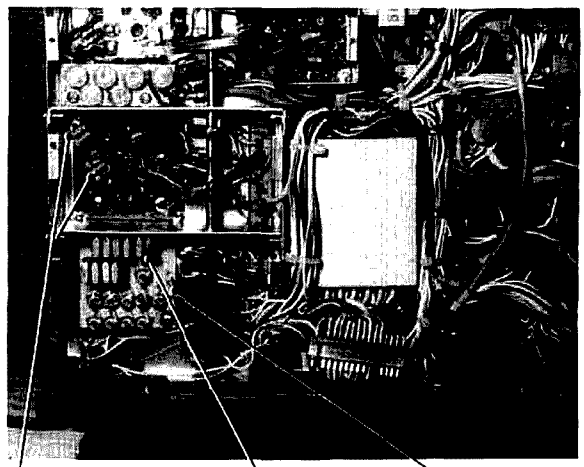
Assuming you have a grid-dip meter available, temporarily tack-solder one of the 45-pF capacitors from the AUX lug on wafer 6 to ground, and, similarly, the other 45-pF capacitor from the AUX lug on wafer 7 to ground. In other words, you'll be shunting the primary and secondary of the filter, each with a 45-pF capacitor. Leave the leads long, and form them open so you can insert the grid-dip meter coil into first the primary, and then the secondary capacitor lead loop.

Power to the FT-301 should be removed. The bandswitch should be in the AUX position. You should find a weak dip on the primary side with the grid-dip meter



In-line cable receptacle which obscures diode-pack installation beneath

Install the 5-diode matrix pack under the in-line cable receptacle.



New crystal, oscillator padder, and padding capacitors across T-1006.

somewhere between 19 and about 25 MHz. You should find a strong dip on the secondary side in the same general range.

With the alignment tool, adjust the core in T-1006 until the strong dip on the secondary side is at 19 MHz. (The weak dip on the primary side should be somewhere between 19 and about 20 MHz; don't worry about the exact value.)

Unsolder the two capacitors from their temporary positions, and, after trimming their leads, solder them permanently at the terminals of T-1006, one shunting the primary and the other shunting the secondary.

If you don't have a grid-dip meter, you may omit the above temporary installation of the capacitors and go immediately to their final installation at the terminals of T-1006. In this case, you won't be able to adjust the core of T-1006 until later.

(The above does not result in an optimal bandpass filter. However, optimal operation over the full range

of 10.0 to 10.5 MHz is not needed. The above approach is more than adequate for the Amateur 30 meter-band and also allows good reception of WWV on 10 MHz.)

You've now completed the conversion of the bandpass filter unit.

step 4: conversion of the tunable filter trimmer unit

The trimmer unit, PB-1446, is located in the center of the transceiver, and the trimmer adjustments are now facing down toward your workbench. We're not going to modify the trimmer unit itself; instead, we're going to augment it with three new trimmers to be located next to the permeability-tuned coils.

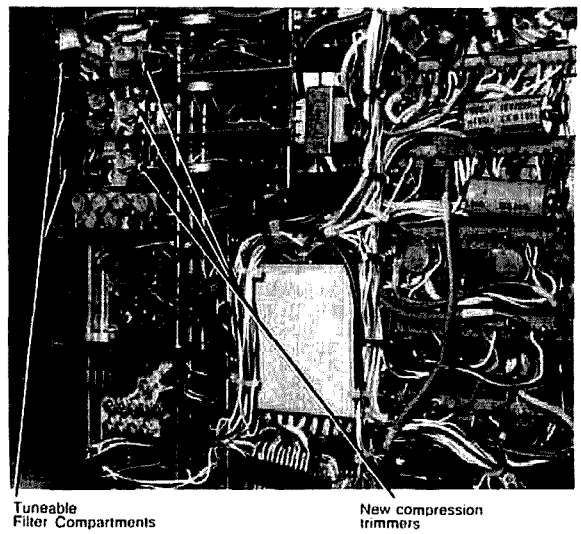
Immediately behind (toward the rear of the rig) the bandpass filter unit of the previous step, you'll find three compartments housing the permeability-tuned coils. The actual permeability-tuning mechanism is facing down toward the workbench. Each of the three compartments contains one (or two) wafers of bandswitch SW2A, and also contains the wires connecting to the underside of the above mentioned trimmer unit.

Using a clean, hot, medium-size soldering iron (about a 60-watt rating is suitable), very quickly unsolder the three flexible braid shields from the three postage-stamp-size PC boards at the base of the permeability-tuned coils, and carefully fold the shield braids back. Also, in the rearmost of the three compartments, unsolder the parallel resistor pack from the same PC board, and carefully fold it back to give full access to the switch wafer.

With a magnifying glass and flashlight, identify the AUX lugs on wafer sections SW2A-9, 10, and 11. (The wafer sections are numbered from the front panel to the rear.) You'll find that the desired lugs are not easily accessible; they're on your left side and are the next-to-lowest lugs. On each of the above wafer sections, the AUX lugs are jumper connected to the lugs for bandswitch positions 10D, 10C, 10B, and 10A. The lug next to the AUX lug is the 10D lug. It's necessary to remove the jumper from the AUX lugs (on just these three wafers, don't remove the jumper on wafer SW2A-8). The jumper wires are wrapped once around the lug.

This is the time for a steady hand; don't attempt this step if you're tired or in a bad mood. It's certainly possible to accomplish the task without damaging anything with the soldering iron. You'll have to dress the various leads to the side a bit to get best access. Plan the job, and practice it once or twice with a cold iron so you can get the feel of the task before you begin.

With the aid of the special soldering tool and a long, thin screwdriver or scratch awl, carefully unsolder and unwrap the jumper wires from the three AUX lugs, one on each of the identified wafers. Don't mangle the lugs. The lug material is thinner gauge than is conventional in American-made switches; treat it with respect. In the two rearmost compartments, you'll have enough room to reach in with a small pair of wire cutters and snip off the unwrapped 1/4-inch (0.635 cm) or so of lead. In the forward compartment, however, such luxury is absent and you'll have to settle for just bending the lead end out of the way, or working it back and forth until it breaks off at the 10D lug.



Tunable filter unit showing new compression trimmer locations.

Prepare three pieces of thin, bare, stranded (flexible) hookup wire, each about 3 inches (7.6 cm) long. Form a suitable hook in the end of each, and carefully solder one of the wires to the now-empty AUX lug on the rearmost wafer, SW2A-11. Bring the free end of the wire up where it is readily accessible. Using the other two wires, do exactly the same thing on the AUX lugs of wafers SW2A-10 and SW2A-9.

The three compression trimmers are to be mounted alike. There's room (just barely) for them to be installed vertically (actually, at about a 60- or 70-degree angle) alongside the compartment shields, one in each of the three compartments, with the adjustment screws facing toward your right, and slightly up. It's necessary to trim the lugs on the compression trimmers so they can fit; the ground end of each trimmer is to be soldered to the postage-stamp-size PC board immediately next to (or under) the corresponding shield braid wire. The other lug on each compression trimmer is connected to the stranded wire you connected to the AUX lug in the same compartment. Be sure to slip a piece of insulation over the bare wire after you've cut it to length so it won't short to the chassis when the compression trimmer is lowered into its final position.

The purpose of the above is to separate the AUX lugs from the 10-meter positions and give each of the three AUX lugs its own compression trimmer to ground.

After the compression trimmers are properly installed, their adjustment screws should be accessible with a long, thin screwdriver or alignment tool. Following installation of the three trimmers and inspection of your work, reinstall the resistor pack lead in the rear compartment and return the three shield braids to their original position.

Now stop and breathe a sigh of relief. The worst is over. The rig will now operate on 30 meters, although **step 5** should be completed before you return the rig to service.

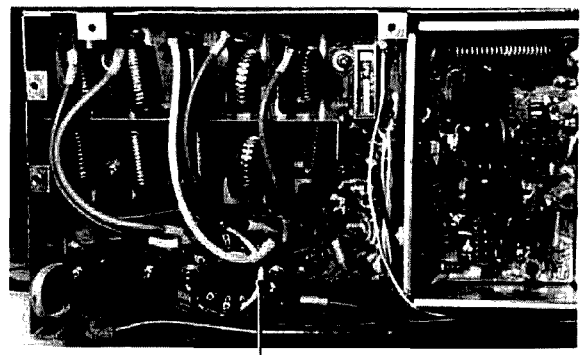
alignment

You may now connect the power cables and turn the transceiver on. Set the band-peaking control on the front panel to a point midway between the 40- and 20-meter band positions. The pointer should be just barely past the numeral 6. Tune the rig to 10.110 MHz, and use the grid-dip meter as a signal source at that frequency. When you've found the grid-dip meter signal, tune the two compression trimmers closest to the front panel for maximum signal. Also peak bandpass filter T-1006 for best reception. If you weren't able to check the adjustment of the crystal tuning earlier, this is a good time to do it.

Connect a dummy load and some kind of power meter to the antenna terminal. Set the mode switch to the CW position and key the rig. Set the drive control so that modest output is obtained and adjust the third compression trimmer (the most rear) for maximum RF output.

Now, observing the RF output, gently peak the second compression trimmer. If the maximum output does not occur at the same compression trimmer adjustment as you obtained for best receiver sensitivity, turn the bahd-peaking control on the front panel slightly toward the 20-meter position and repeat the alignment of the two front-most compression trimmers on receive. Then check the adjustment for maximum transmitter power as above. It should not take more than one or two tries to find a satisfactory set of adjustments.

At this point you should have no trouble hearing 30-meter signals on the air, and, at the right time of day, a very strong WWV on 10 MHz. You might find the receiver gain to be a little lower on this band than on the other bands. Much of the reason is because of the 9-MHz trap in the antenna lead. If you wish, simply short it out or tune it somewhat lower than 9 MHz and accept the slight compromise with 9-MHz suppression. (The 9-MHz trap is T-1401, and is located on trimmer unit PB-1446.) In any case, the receiver gain is entirely adequate for everyday use.



One of the two new jumpers Connecting 20 meter LPF to the AUX terminals

Low-pass output filter assembly on rear apron.

You may find it easy to overdrive the final, and that the drive control must be set back to a lower level than customary. The reason is that the exciter drive-level potentiometer serving the 10-meter band also serves the 30-meter band now. Readjust it if you wish, following the instructions for the VR Unit PB-1448 found in the instruction manual. It is easier, in my opinion, to simply reduce the drive level using the front panel control. You're now ready for the final conversion step.

step 5: conversion of the output low-pass filter unit

In the original design of the FT-301, Yaesu didn't include any low-pass filtering for the 10- and 11-meter bands. Even though the power amplifier is push-pull and is essentially linear, it's good practice to use a filter on the output. There is not sufficient room to include a low-pass filter specially dedicated to the new 30-meter band position.

A practical alternative that works well for me is using the 20-meter low-pass filter for 30 meters as well as for 20 meters. With this arrangement, no second harmonic was detectable at a listening location about 3/4 mile (1.2 km) from my location. Higher harmonics should be suppressed by the 20-meter filter, which is a conventional constant k design, without m-derived notch sections.

Remove the power cables and turn the transceiver so the rear apron faces forward. Remove the power amplifier (assuming you do not have the QRP version). The low-pass filters are located in the compartment along the rear apron. With the power amplifier (or the rear cover, for the QRP model) removed, both filters and the wafers of switch SW2B are visible. It is necessary to get at both wafers. Proceed as follows:

Locate the shaft coupling that couples the rear end of the shaft of SW2A to the forward end of the shaft of SW2B. Loosen the two rear set screws. Be sure you understand what position the switch is left in, and

which lug on the SW2B wafers correspond to AUX and to 20-meter positions.

There are four screws holding the PC board with the filter coils and four screws holding the mounting plate for SW2B. Remove all eight. *Be especially cautious removing the latter four*; it's easy to distort the switch lugs, and once bent out of shape it is a terrifying job to repair them. Here's where you'll need caution and well-fitting screwdrivers.

With the eight screws removed, lift the PC board and switch body together as a loose assembly, gaining access to the lugs on both wafers.

As before, remove the jumpers that connect the AUX lugs to the 10D lugs on both wafers. Solder a new jumper of fairly heavy wire connecting the AUX lug on the top wafer to the 20-meter lug on the same wafer. Do the same on the bottom wafer. The purpose is to use the 20-meter filter on 30 meters as well as on 20 meters.

After inspecting your work, reassemble the filter assembly, remembering to have SW2B in the correct position with respect to SW2A before tightening the shaft coupling. To ensure proper mechanical alignment, tighten the shaft coupling while holding SW2B firmly against its mounting surface, before you tighten the four switch plate mounting screws.

To avoid losing the four switch mounting screws, position the transceiver vertically with the front panel facing down. (Two old copies of the Callbook are just the right size for supporting the edges of the front panel so that no mechanical stress is placed on the knobs.)

With reassembly of the low-pass filter unit and replacement of the rear panel/power amplifier, you've completed the conversion.

final cleanup

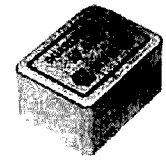
Reassembly is easy, and follows the reverse order as disassembly. You'll want to make one final alignment before replacing the shield plate. The final alignment should be done at about 10.125 MHz and should follow the same procedure as above. The receiver alignment should be done while listening to a weak signal on the air. Adjust T-1006 this time for maximum RF drive in the transmit position; doing this makes it a little easier to tune it to the peak position.

You may notice a large amount of 1-MHz marker leakthrough on 9 MHz while working on the receiver. With the shields replaced and the rig buttoned up, this will no longer be a problem.

No special instructions are needed to enjoy 30 meters. The rig operates on this band just as if it came factory equipped with the capability. Signal reports on the air have been good, and I enjoy having WWV on 10 MHz conveniently at hand.

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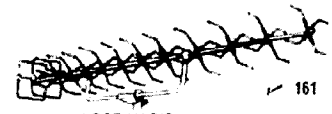
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PRACTICALLY SPEAKING ... Joe Carr K4IPV

build a bench power supply

A popular project for Amateurs and other electronic hobbyists is the direct current (DC) power supply for the shop workbench. Amateurs are often tempted to build their own power supplies because commercial models may be relatively expensive, especially when "junkbox" parts are readily available. We also often find that voltage and current rating, in particular, are not to our liking. For example, many ±12 VDC power supplies offer only 100 or 500 mA of current — which seems terribly little!

The DC power supply described below can be easily built by almost anyone, is easily adapted to other voltages or currents and as shown, is already more flexible than most commercial bench power supplies.

In designing this power supply, my selections were dictated by my own needs, which are, I hope, similar to yours. Because I work a lot with operational and other linear IC amplifiers,

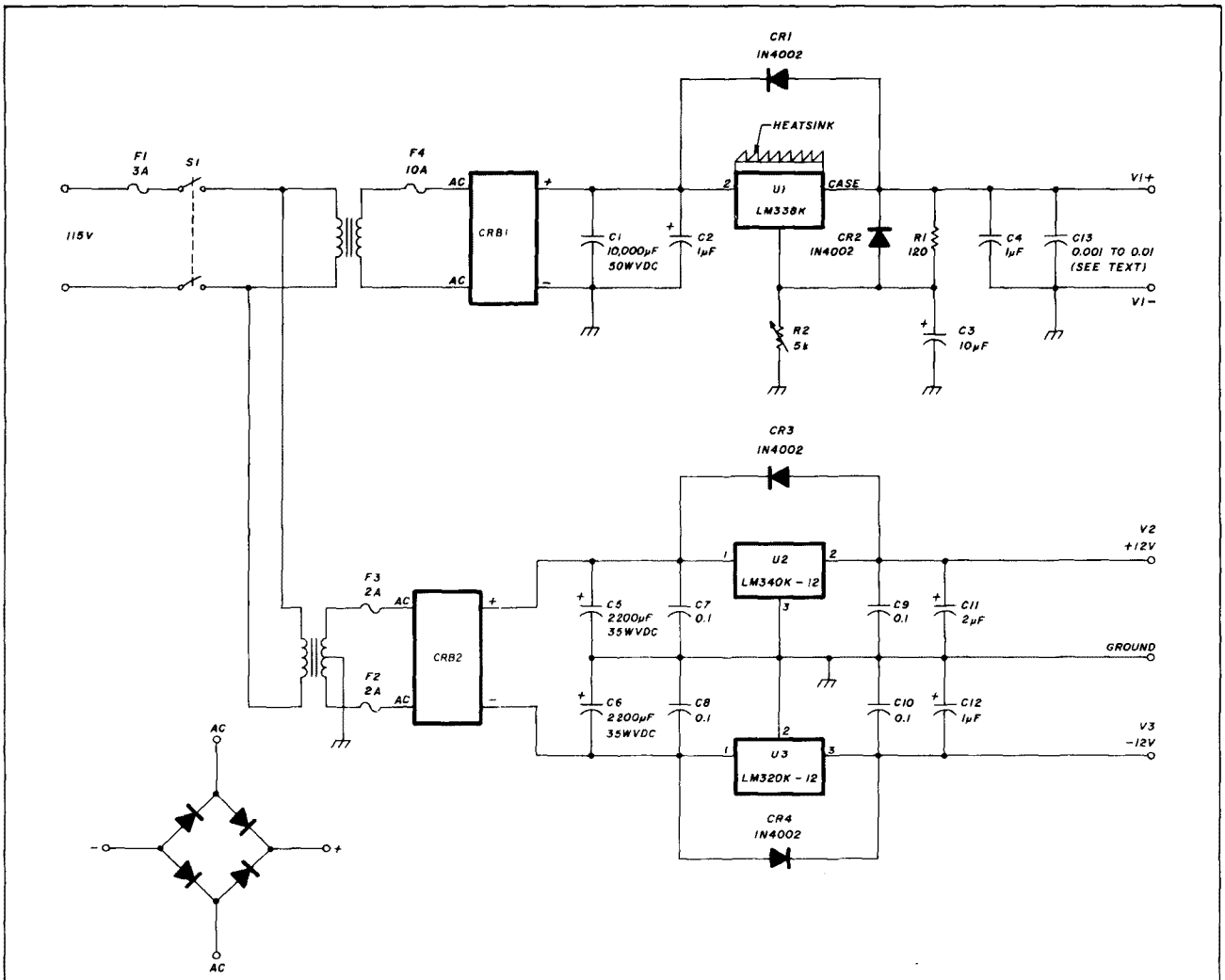


fig. 1. Circuit diagram of multivoltage adjustable power supply.

I needed either ± 12 VDC or ± 15 VDC at a current of 500 mA or more per voltage (e.g., – 12 VDC at 500 mA, or more). Because some of my projects have been extensive, and also because 1-ampere components are so easily available, I built a ±12 VDC dual-polarity power supply that provides 1 ampere at both voltages.

The DC power supply also had to provide a variable 0 to 15 VDC at 5 amperes. In addition, a supply capable of being adjusted for 5 volts would power TTL digital circuits that include most of today's 64K single-board microcomputers.

Digital projects are not my only reasons for needing a brutish variable voltage power supply. Sometimes I repair (or operate) my 2-meter FM mobile rig in the house. And others will occasionally ask me to look at, test, fix or otherwise wave a magic wand over VHF-FM marine radios — some of which are merely high-priced versions of popular ham rigs! For these jobs, I set the output voltage to + 13.8 VDC (or + 13.6 VDC for some models), and can deliver up to 68 watts with the 5 amperes of current available. This power supply will handle a 50-watt mobile rig nicely.

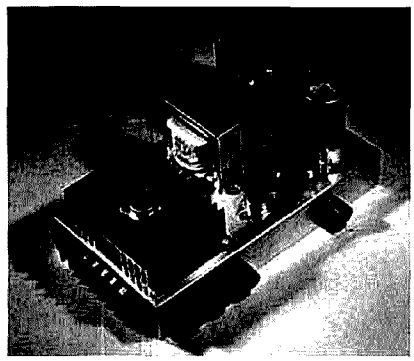


fig. 2. Completed power supply.

Another design requirement was that all voltages be regulated. This requirement not only allows precise operation, but also reduces rectifier ripple to nearly zero. While most of the projects I build don't specifically require voltage regulation, none are harmed by it. On the other hand, projects that require voltage regulation usually can't do without it, so I provide regulated voltages to all projects.

In the sections below, we'll discuss the design of the power supply, paying some attention to the rules of thumb used. These rules will permit you to design and build similar power supplies of your own. After all, the circuit for the + 12 VDC fixed supply is good for any 100 mA to the 3-Ampere, three-terminal IC voltage regulator.

parts list

T1	*12.6 VAC RMS at 5 amperes (or more!)*
T2	*25.6 VAC RMS at 3 amperes center-tapped*
CRB1	*50-volt PIV at 12 amperes bridge rectifier*
CRB2	*50 volt PIV at 1 ampere bridge rectifier (note: 1000 volt PIV used in the original).*
C1	*10,000 μF/35 t0 50 WVDC (or two 5000 μF in parallel)*
C2	*1 μF/35 to 50 WVDC*
C3	*10 μF/25 WVDC*
C4	*1 μF/25 WVDC*
C5,C6	*2200 μF/35 WVDC*
C7,C8,C9,C10	*0.1 μF/50 WVDC*
C11,C12.	*1 μF/25 WVDC*
C13	*0.001 μF to 0.01μ/50WVDC disc ceramic*
CR1 — 4	*1N4002 (or any 1N4XXX up to 1N4007)*
R1	*120 ohms/0.5 watt*
R2	*5000 ohm potentiometer*
S1	*SPDT switch rated for 110 VAC or higher*
U1	*LM338K*
U2	*LM-340K-12 (or "K" packaged 7812) LM340K-12*
U3	*LM320K-12 (or "K" packaged 7912)*
F1	*3-ampere fuse*
F2,F3	*2-ampere fuse*
F4	*10-ampere fuse*

Miscellaneous: four chassis-mounted fuseholders, heatsink for U1 (see fig. 2), chassis or cabinet, output terminals of your selection (most will use binding posts or barrier strips).

design

The power supply has two sections, as evidenced by the 1.0- to-15 VDC at 5 ampere variable and ± 12 VDC at 1 ampere (per side) dual polarity fixed supplies. Because custom transformers are neither cheap nor easily obtained, I used separate transformers for this project.

The requirement for most three-terminal IC voltage regulators is that the input voltage be + 2.5 volts higher than the rated output voltage. For the 12 VDC supplies, therefore, we need 14.5 VDC or more from the rectifier/filter circuit; for the 0 to 15 VDC supply, we need 17.5 VDC from the rectifier/filter.

So how do we obtain the needed voltages? In most cases, we will have to specify the AC secondary voltage of the transformers, so we must be able to relate that specification (an RMS voltage) to the required voltage. Keep in mind that for sine waves, the output of the rectifier is very nearly the peak voltage of the transformer secondary winding. In order to find the minimum RMS voltage rating, we must multiply the required DC voltage needed by 0.707. For the 12 VDC supplies, where the input must be 14.5

VDC or more, then: 0.707 × 14.5 = 10.5 volts RMS, which means that the RMS rating of the transformer secondary must be 10.5 VAC RMS (or more). For the 0 to 15 VDC supply, where the voltage to the regulator input must be 17.5 volts DC or more: 0.707 × 17.5 = 12.4 VAC RMS (or more). In both cases, the next higher "standard" transformer rating is 12.6 VAC RMS.

The transformer used for the 0 to 15 VDC at 5-ampere supply needs a secondary current of 5 amperes or more. Being conservative in such matters (by virtue having had to repair too many burned-out power supplies over the past two decades), I selected a 12.6 VAC RMS at 8 ampere model from a mail-order catalog, though I would have preferred 10 amperes). Note: surplus military transformers are usually derated considerably, so they can generally be used by Amateurs at a higher rating.

The ±VDC dual polarity supply presents a special problem because we would ordinarily need two 12.6 VAC transformers, one for each polarity. For this case, I selected a 25.2 VAC RMS (or higher) transformer that has a center-tapped secondary winding. With the center tap used as the ground (or common) reference, we essentially have two 12.6 VAC RMS transformers sharing a common secondary.

The current rating of the 25.2 VAC RMS transformer needs to be at least 2 amperes (1 ampere for each polarity), plus a safety margin. This extra rating is needed because I used a single fullwave bridge rectifier as if it were two halfwave bridge rectifiers. This rectification scheme requires a 40 percent margin, or 0.8 amperes in this case. Thus, the transformer rating had to be 25.2 VAC RMS (or higher) at 2.8 amperes (or more). My local Radio Shack had one on sale (but not in the catalog!) that was rated at 25.6 VAC RMS at 3 amperes with a center-tapped secondary. It filled the bill nicely.

Rectifiers CRB1 and CRB2 were selected according to two different rating criteria: peak inverse voltage (PIV) and forward current. Although both rectifiers were bought as pre-made bridge "stacks," yours can be made from discrete rectifier diodes as shown in the inset in **fig. 1**.

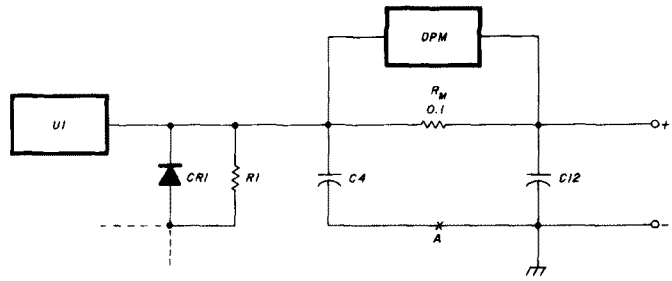


fig. 3. Using a DPM for output current metering.

The required PIV voltage rating is determined from the applied forward voltage, and is a minimum rating; higher PIV is permitted, and may even be desireable. The usual rule of thumb is that the PIV rating of the diode in a filtered DC power supply be 2.82 times the applied RMS forward voltage. Since 12.6 VAC RMS is applied, the minimum acceptable PIV rating is 2.82 × 12.6 VAC RMS, or 36 volts. The next higher standard PIV rating is 50 volts PIV.

The forward current for CRB1 is 5 amperes; and for CRB2 it is 1 ampere. Because I prefer a substantial margin, I selected a 12-ampere chassis-mounted model for CRB1. Rectifier CRB2 was rated at 1 ampere at 1000 volts PIV because these are almost universally available, and stacks greater than 1 ampere are both more expensive and less easily obtained.

The filter capacitors are selected according to a 2000 μF/ampere rule of thumb, so 10,000 μF was used for C1 and 2000 μF was used in each of the 12 VDC supplies (C7 and C8). In both cases, the voltage rating (WVDC) must be greater than the rectified output voltage — or not less than 25 WVDC given that standard voltages must be used. For a margin of safety, I selected 50 WVDC for C1 and 35 WVDC for C7 and C8.

The other capacitors in the circuit are for noise immunity protection. The lone exception is the optional capacitor, C13. This capacitor is needed by those who work on transmitters or other devices that emit RF energy. Radio frequency signals that get onto the output leads are transferred into the power supply and become a source of problems. Place a 0.001 μF to 0.01 μF disk ceramic capacitor across the output. This capacitor must be mounted directly on the output terminals, not on the circuit board. In general, use a value closer to 0.001 μF if you work only on VHF equipment, and closer to 0.01 μF if HF and lower frequencies are anticipated. Keep in mind that the HF value might be indicated if you plan to monitor 2 meters while operating on HF, even though the power supply will never be used for HF equipment.

There are four 1N4002 (or any 1N4XXX-series diode up to 1N4007) rectifier diodes in this circuit. These diodes are used to protect the IC voltage regulator, especially at turn-off. Charge stored in the various capacitors can be dumped back into the regulators in a disastrous manner, and the diodes provide a less dangerous path. Normally, diodes CR1 through CR4 are reverse-biased.

construction

Figure 2 shows the construction of the power supply. Although I used a simple aluminum chassis and bottom plate, you might opt for a cabinet. My

selection was dictated by the physical configuration of my workbench.

Both transformers are mounted on the chassis, as are capacitor C1 and the heatsink for U1. Keep in mind that either U1 must be insulated from the heatsink (the method I selected), or the heatsink must be insulated from the chassis. The reason for this requirement is that the output terminal of U1 is the regulator case. If you opt for insulating U1, then buy a transistor mounting kit.

Another option is the type of potentiometer selected for R2. In my case, I used a remotely located potentiometer on top of the workbench; you might prefer the pot on the same cabinet as the rest of the supply. Alternatively, you might prefer a trimmer potentiometer for set-and-forget operation.

There are two other options for you to consider. First, do you want an output voltmeter, especially on the 0 to 15 VDC supply? I used an outboard meter — my normal bench digital multimeter. You, however, may want to obtain either analog or digital panel meters and install them across the 0 to 15 VDC output.

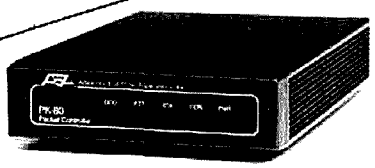
The second option is a current meter for the 0 to 15 VDC supply. A 0 to 5 ampere analog meter can be connected in series with either output line, but if you want a fancy digital meter, take a look at **fig. 3**, in which the circuit of the 0 to 15 VDC supply is shown modified for metering. We can either break the positive output line as shown, or break the negative output line at point "A" and insert a 0.1 ohm, 2.5 watt resistor. These resistors can be made from several parallel connected wirewound resistors, or bought as a single 0.1-ohm unit. (Automotive electronics use such resistors under the name "fusistors."). This value is selected because it will produce a voltage drop (according to Ohm's law) of $E = IR = (1\ Amp) \times (0.1\ ohm)$ or 0.1 volt per ampere. This means that the output current is numerically the same as the voltage reading. For example, when 5 amperes are drawn, the voltage drop across R_M is 0.5 volts, or 500 millivolts. If you buy or build a 0 to 1.999 volt 3-1/2 digit digital panel meter (DPM), then the reading at 5 amperes would be "5.00." Thus, we can monitor current down to steps of 10mA (example: "2.45 V" corresponds to 2.45 amperes). Turn off or otherwise delete the "V" label on the meter, however.

conclusion

The project above is a nearly ideal bench power supply for almost all electronic hobbyists — hams, computer enthusiasts and general electronic hobbyists alike. It will provide voltages up to 15 VDC at currents up to 5 amperes and also be set to +5.00 volts to power digital projects up to 5 amperes. The dual polarity is more than enough for all but the most demanding (and large) projects. Yet it's duck soup to build!

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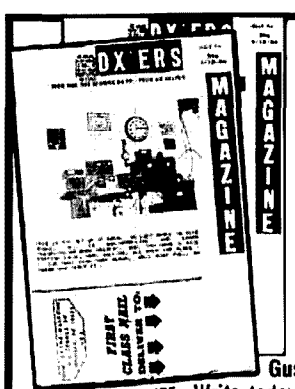
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ham radio TECHNIQUES

Bill Orr W6SAI

more on the G5RV antenna

Although the G5RV multiband wire antenna has been popular overseas for some years, it's not well known in this country. But I've written about it occasionally in this column, and from the feedback I get in my mail, interest in this antenna appears to be growing. And why not? It's an inexpensive antenna that allows operation on more than one Amateur band without the necessity of using an auxiliary antenna tuner.

The basic antenna, popularized some years ago by R.L. Varney, G5RV, is shown in **fig. 1**. The design covers 7, 14, and 28 MHz, supposedly with a low value of SWR on each band, and is fed with a 75-ohm coax line. In practice, the measured SWR on the line is below 1.5:1 at the low frequency ends of the 10- and 20-meter bands, but is quite high across the 40 meter band.

Since most ham equipment is standardized on a 50-ohm transmission line, the G5RV design provides less than optimum performance when this coax is used for the feeder.

In 1981, W5ANB designed a modified version of the G5RV antenna[2] that provided good performance on the 7, 18, and 28 MHz bands when fed with a 50-ohm transmission line (**fig. 2**). Because the 18 MHz band is, as of this writing, not yet open to Amateur communications, the W5ANB antenna is useful only on the 7- and 28-MHz bands. Even so, this design demonstrates that changing the antenna dimensions moves the SWR response curve between various Amateur bands.

There the matter rested until Brian Austin, ZS6BKW, derived a computer program for the G5RV antenna that permits the designer to vary the antenna parameters and observe the results. The starting point of his investigation was the G5RV and the W5ANB models. Brian experimented with various modifications of these

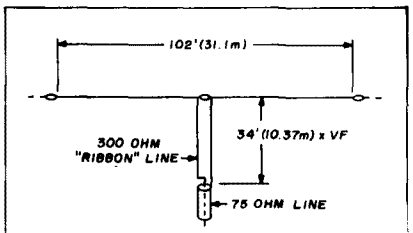


fig. 1. The original G5RV multiband antenna.

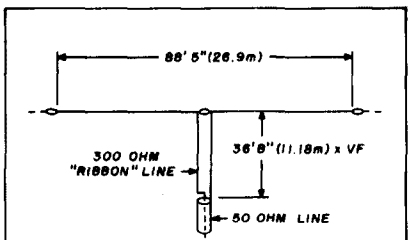


fig. 2. The W5ANB version of the G5RV antenna.

designs and tested his results on a model antenna range. The experiments eventually provided a G5RV-type antenna that produced a better impedance match to a 50-ohm line on various bands than did either of the two earlier designs.

the ZS6BKW multiband antenna design

Dimensions for the ZS6BKW antenna design are shown in **fig. 3**. The antenna is fed with a 50-ohm coax line and provides a good match to it on the 7, 14, 18, 24, and 28 MHz bands. Only the 21 MHz band is omitted. The respective SWR curves for this antenna are shown in **fig. 4**.

This design looks like the long-sought-after multiband antenna, but note that the two-wire transformer section must be built using 400-ohm line. Unfortunately, this line is not commercially available, although a home-made line can be built from No. 14 gauge wire spaced 0.9 inch (22.8 millimeters) between centers. Alternatively, a commercial 450-ohm ladder line with a polyethelene dielectric may be used. This oversized line has holes punched in the insulating web and the impedance is close enough to the desired value to provide satisfactory operation.*

Although the design doesn't call for it, I believe it's a good idea to a place 1:1 balun at the junction of the coax line and the two-wire transformer section. Alternatively, the coax can be wound into an RF choke coil to suppress unwanted antenna currents from travelling down the outside of the outer shield of the line. Four turns of line wound into a coil whose diameter is about 15 times the diameter of the cable will suffice.

the ZS6BKW antenna using 300-ohm line

Brian, ZS6BKW, determined from his computer program that by changing the lengths of the antenna sections, the design could be modified to use a 300 ohm line for the two-wire transformer. This line is universally

*Saxton No. 1562 line (450 ohms), manufactured by Saxton Products, 215 North Route 303, Congers, New York 10920.

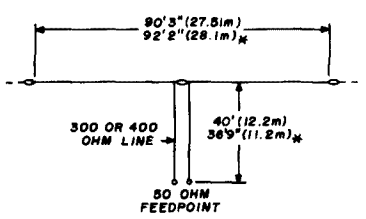


fig. 3. Physical dimensions for ZS6BKW multiband antenna. For 300 ohms line, use lower dimensions (*).

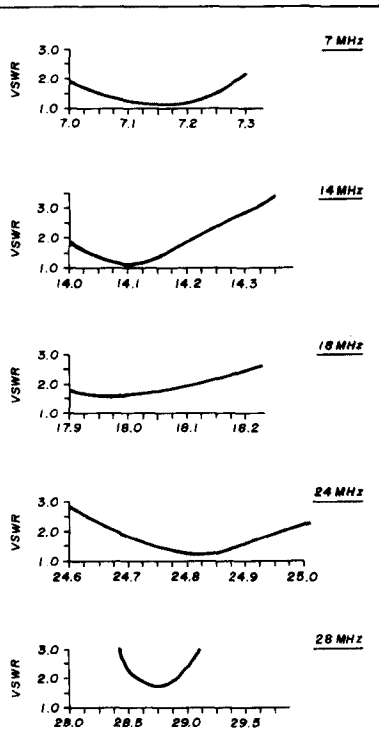


fig. 4. Representative SWR curves of ZS6BKW antenna.

available. The SWR curves for the various bands for this antenna are very close to those shown in **fig. 4**. This modified version seems to be the inexpensive, five-band wire antenna desired for many years by radio Amateurs!

building the ZS6BKW multiband antenna

A complete set of physical dimensions for the ZS6BKW antenna is given in **fig. 3**. It's easy to construct. Just string the flat top out between two supports and attach the transmission line so that it departs at right angles to the antenna. No special precautions need be observed beyond simply waterproofing the joint between the two-wire line and the coax line to prevent moisture from entering the coax.

The SWR curves on each band are affected by the height of the antenna above ground. If you want to alter the shape of the SWR curve on a particular band, you can change the antenna's elevation — but remember that if you do this, the SWR curves for the other bands will also change. The best SWR curves for all bands were found at an antenna elevation of about 42 feet (13 meters) above ground, although good results were obtained with the antenna as low as 23 feet (7 meters).

Tests were run on the ZS6BKW antenna configured as an inverted-V. The apex of the V was 40 feet (12 meters) high. When the angle between the wires was lowered from 180 degrees to 90 degrees, it was found that the resonant frequency dropped about 80 kHz in the 14 MHz band and about 125 kHz in the 24 and 28 MHz bands. This was thought to be due to the additional capacitance between the ends of the antenna and ground.

the RF light bulb revisited

In my October, 1985, column, I spoke about the QRM-possibilities of the RF-actuated light bulb and voiced the hope that "some enterprising Radio Amateurs would have appropriate facilities at hand to examine RF light bulbs."

The investigation had been taken up by the ARRL, which ran extensive tests on the bulbs and filed comments with the FCC on Docket 83-806, which concerns the regulations that may be required for these devices.

Paul Rinaldo, W4RI, Editor of *QST*, forwarded a copy of the ARRL filing to me. The extensive tests indicated that with a typical American Phillips SL-18 lamp, " . . . interference may be expected in residential environments from RF lighting devices to broadcast band receivers and, to varying extents, HF receivers when the antenna is located near the bulb. However, the RF noise emanating from the test lamp was insufficient to cause interference to a typical Amateur Radio station with outside antennas if the lamp is inside or located in an adjoining residence."

The report continues, "From the point of view of interference potential to Amateur Radio stations, the League believes it both necessary and sufficient for the Commission to require labelling of each RF light bulb to provide user information to educate consumers and users about interference potential to radio receivers . . . . This is important not only from the point of view of the licensed Amateur who might purchase a bulb for his own use but also to insure that the neighbor of a licensed Amateur who suffers interference to a broadcast receiver from an RF light bulb does not wrongfully blame the Amateur for "causing" the interference."

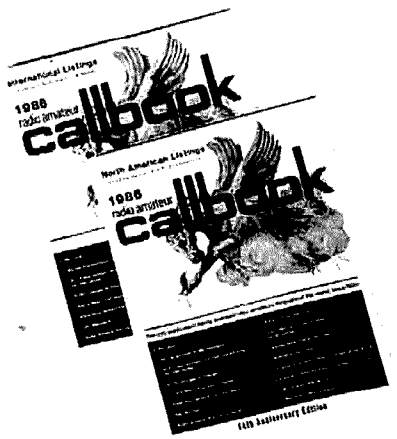
good show, ARRL!

However, the battle over the RF light bulb continues. In the fall of 1985, the FCC relaxed certain regulations governing the use, marketing, and certification of RF lighting devices and ISM (industrial-scientific-medical) devices, all of which broadcasters claim will interfere with AM reception.[3] National Association of Broadcasters (NAB) Staff Engineer Michael Rau stated that the relaxed rules makes AM reception "especially vulnerable" because they do not address interference experienced below 30MHz.

A battle seems to be brewing between the NAB, which represents broadcasters, and the NEMA (National Electrical Manufacturers Association), which states that the FCC is being "very realistic in its simplification of administrative responsibilities and certification laws involving the marketing of RF lighting devices."

So there you are. The NAB and the NEMA are both pulling and hauling on the FCC with regard to regulations

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concerning certification of ISM equipment and the residential use of consumer RF devices that might cause radio interference. The outcome of this interplay of forces will probably affect Amateur Radio in the future.

CW and all that

I've noticed a great improvement in CW operation on the 20-meter band during the past few decades. The electronic key and the keyboard have produced some high-class signals whizzing along at 30 to 40 WPM. In general, the average code speed has increased greatly over that used by Amateurs in the past.

When I first operated 20 meters in the late 1930's, there were a few speed merchants on the air, but most Amateurs plodded at 8 to 10 WPM. Why were communications so slow in those bygone days? Because Amateur Radio was much less sophisticated then—and a good bug key cost the equivalent of two weeks' wages! The hand key was king.

But code champions existed then, too. Code speed contests at the larger Amateur Radio conventions were popular.

Copying high speed code by pencil or typewriter was an art, as the following story of a high speed CW contest as seen through the eyes of one contestant, taken from the November, 1933, issue of *QST*, suggests. The contest started out at 8 WPM and the code speed was gradually increased by 5 WPM each minute until only one contestant was left . . . .

At 8 words a minute, you sit back and twiddle your thumbs, you yawn, and wish to heaven that the "lid" would get off the air. At 15, you take up your pencil and leisurely jot the stuff down on a piece of paper. At 20, you see the first signs of life. For a minute or two you sit back and copy, and then, on second thought, you hitch your chair forward a bit and straighten the paper. At 25, you quit "laying behind"; you decide to close the gap until you are about a word behind the sender. Not so bad, now. At 30, the fun begins. You can read it all right, but the pencil seems to be getting a little sluggish – better make a grab for a "mill." At 35, you begin for the first time to think about errors: "How many am I allowed on a 5-minutes' run of this? At 40, it gets hotter, and darn suddenly, too. The last 5 words a minute have more mustard on them, it seems, than the first 30. You are holding your own with many a crack commercial radio or telegraph operator now. You quit worrying about single wrong letters and start hoping you can put a typewritten line down without leaving a word out. At 45, the jig is up. You quit, but half a dozen of the chaps go on. You make a grab for the transcription (if a judge) and by dint of much mental agility you may contrive to "follow" the transmission, but you are glad you are no longer trying to put it down on the typewriter. At 50 words a minute, the dots and dashes get blurred and jumbled. The letter O, instead of being designated by three ordinary dashes, comes through as one dash, and that one exceptionally staccato. The common "and's" and "the's" whiz by undeciphered. At 55, it is just a lot of static – no sense now in trying to hear anything . . . .

todays speed merchants

It seems that 40- or 50-WPM QSOs can be heard daily on most of the Amateur bands. The keybord and the code reader have made great strides for Amateur communication. But how many hams can copy 45 WPM on a "mill?" I know a few that can, and there must be others. I also know a handful of old-time Amateurs who can copy up to 100 WPM in their head! Now, that's CW ability!

What is the world's CW speed record today? I don't believe I know. Does anybody out there know?

references

1. R.L. Varney, "An Effective Multiband Aerial of Simple Construction," *RSGB Bulletin*, July, 1958, pages 19-20.
2. T. Nicholson, "Compact Multiband Antennas Without Traps," *QST*, November, 1981, pages 26-27.
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ham radio

VHF/UHF WORLD

Joe Reisert W1JR

reflector antennas: part II

Last month's column dealt primarily with the electrical and mechanical considerations for designing a parabolic dish type of antenna. Various design tradeoffs were discussed, especially in regard to choosing the proper f/d (focal length to diameter ratio).[1]

This month's column will continue in the same vein, with last month's material to be repeated only as required. Emphasis will instead be placed on additional considerations, recommended feed systems, and some typical construction techniques. You should then be able to build your own parabolic dish or design a feed system for an existing dish and place it in service.

other parameters

Last month's column emphasized the f/d ratio and the comparatively low efficiency 50-60 percent of the operating parabolic dish antenna, but the apparent disparity between 50-60 percent and the efficiencies shown in **table 2** were not explained. The reason for the disparity is that there are other factors that must be considered when calculating efficiency. So as not to be chided for passing over the subject too lightly, I'll briefly discuss some of these parameters here.

As discussed last month, aperture illumination and spillover are not the only factors that reduce the efficiency of a parabolic dish type of antenna. Other losses include but are not limited to structural imperfections, reflector surface accuracy, reflector porosity, polarization errors, and feed system blockage, among others.

It should go without saying that the shape of the reflector must follow a parabolic taper throughout its surface. Any departure from a true parabolic curve causes random phase errors which decrease performance. This includes the supporting ribs and the reflector surface tolerance and has been covered extensively in the literature.[2]

It should also be mentioned that the required accuracy of the parabolic curve is a function of the f/d ratio of this dish; f/d ratios greater than 0.45 require tighter tolerances than those under 0.45. As rule of thumb, the mechanical tolerances of the reflector should be held to 0.05 wavelength, with a maximum of 0.1 wavelength at the highest frequency of operation. This suggests a tolerance of 0.5 inches (1.3 cm) for the 1296 MHz (23 cm) band. Typically the reflector tolerances for a well constructed reflector should add up to a loss of only about 0.2 dB.

The reflector has another form of loss due to the porosity of the surface. If the reflector isn't made of solid material, some of the energy imping-

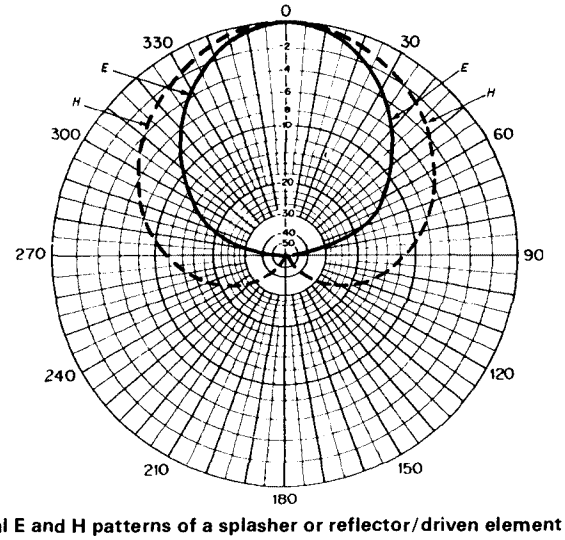


fig. 1. Typical E and H patterns of a splasher or reflector/driven element type feed as described in the text. Note the considerable difference between each plane's beamwidth.

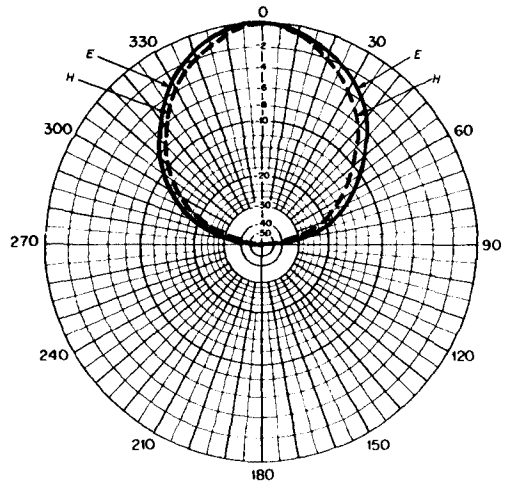


fig. 2. This is the approximate pattern of the EIA feed system as described in the text. Note that the E and H planes are clean and have similar beamwidths.

ing on the surface will pass through it. This is a function of the largest dimension of the holes and the diameter of the wire in the screen. As a rule of thumb, the losses due to a typical mesh type of reflector will be 0.3 to 0.5 dB when the largest dimension across the mesh is about 0.1 wavelength at the frequency of operation.[3] Therefore, 1 inch (2.54 cm) diameter chicken wire screen should not be used above 23 cm.

Polarization errors are caused by the type of feed and the f/d ratio of the dish. Lower f/d ratio (0.25 to 0.4) designs have greater cross polarization losses than higher f/d ratio (0.45 to 0.6) dishes. The additional loss due to polarization is typically 0.15 to 0.2 dB.

Another factor to be considered is aperture blockage, which is affected by the size of the feed as well as the feed holding structure. Intuitively, the larger the feed structure, the larger the aperture blockage and the lower the gain. Often Amateurs fail to recognize that the feed supporting structure also contributes to aperture blockage.

The total aperture blockage can be determined by adding up the cross sectional area of the feed and supporting structure and comparing it to the overall surface area of the dish using the following equation[3]:

$$L = 10 \log \left[1 - 2 \left(\frac{B}{A} \right) \right]^2$$

where L is loss in dB, *B* is the aperture blockage area and *A* is the area of the dish with both *A* and *B* in the same unit of measure. For example, a 20-foot (6.1-meter) diameter dish has an area of 314 square feet (29.2 square meters). If the total aperture blockage due to feed and supports is 5.75 square feet (0.534 square meters), the degradation will be approximately 0.32 dB.

The losses due to these factors will be approximately 1 to 2 dB, depending upon how well the dish is designed and built. Adding this to the approximately 1 dB of illumination taper discussed in reference 1 yields a 2 to 3 dB typical loss versus *the total aperture of* the dish. Therefore a typically well constructed parabolic dish has only 50 to 60 percent efficiency based on its physical aperture.[4]

feed systems

Ironically, the feed system becomes the most important item in the design of a parabolic dish type of antenna. This is true even if you can build a perfect reflector with no deviation from a true parabolic taper, since you still must illuminate the dish properly for maximum efficiency.

Let's reiterate what's required of the feed system. First, it must provide the proper taper at the rim of the dish. Hence the RF level of the feed must be down approximately 10 dB at the rim of the dish from the level at the center of the dish.[5] Also remember that the space attenuation at the rim of the dish, because of the angle off boresight, increases the feed pattern taper and must be considered as shown in fig. 7 of reference 1.

Next, the feed system should have a clean lobe-free pattern that's relatively symmetrical in both E and H planes. A 10 percent difference between E and H beamwidth is acceptable. Also remember that most feeds are not circularly symmetrical, even though the parabolic dish is! Hence, there's further mismatch. If you desire to use circular polarization for OSCAR or EME operation, the feed design is even more critical.

A simple feed pattern is shown in fig. 1. This is the pattern for a typical dipole with a reflector that's often referred to as a "splasher" or dipole-disc feed, which used to be one of the most popular feed systems.

Note that the beamwidths in the E and H planes are quite different. Since the beamwidths are so dissimilar, either the E or H plane can be optimized for the particular dish at the expense of the other plane. Hence it is not a recommended feed system for a parabolic dish type of antenna, except where performance is not critical.

Figure 2 shows the pattern of the EIA feed.[6] Note that it has a clean pattern as well as similar E and H beamwidths. This is the ideal combination for an efficient feed system.

Finally, you have to know the taper of the feed in order to place the proper attenuation point at the rim of the dish. Figure 3 includes a handy graph for doing this. Using fig. 3 you can estimate the attenuation off boresight if you know the −10 dB beamwidth. If you know the beamwidth at another level, it can be normalized to the graph. For example, if the half power (−3 dB) beamwidth of your feed is

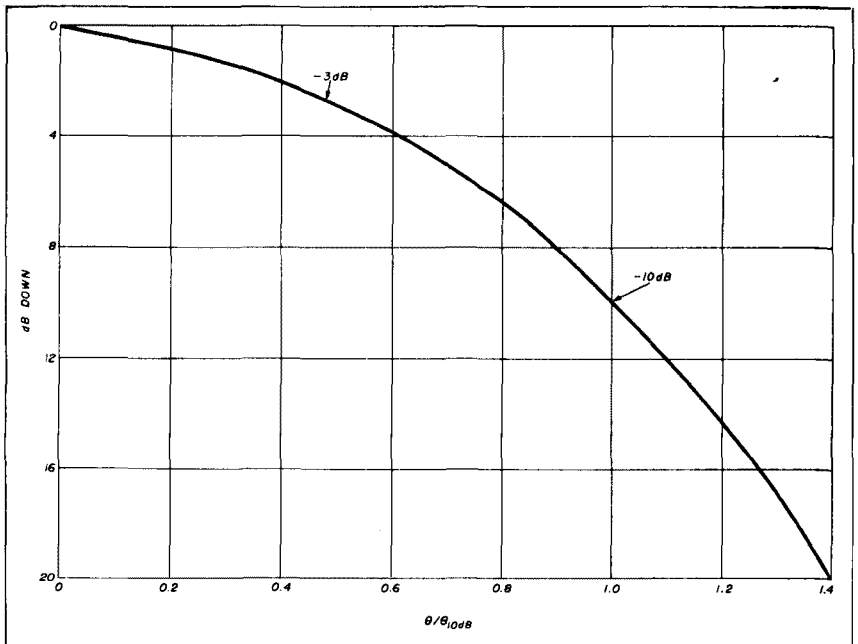


fig. 3. This graph represents a universal pattern which can be used to estimate the beamwidth of a typical feed pattern if the beamwidth is known at 10 dB below the main beam. If the beamwidth is known at another level, it can be normalized to the pattern as explained in the text.

60 degrees, the −10 dB beamwidth will be approximately 1.92 times this value (1/0.52) or 115 degrees. Conversely, if the −10 dB beamwidth is 120 degrees, the −3 dB beamwidth will be 0.52 times 120 degrees, or 62.4 degrees.

Let's discuss some typical parabolic dish feeds often used by Amateurs. They are (but are not limited to): the "splasher," the EIA reference antenna, and the rectangular and circular waveguides.

The dipole in front of a disc reflector shown in **fig. 4A** has been around about as long as any type of feed as discussed above.[5] It's relatively simple to construct, has a built-in balun, and has a relatively small reflector, causing relatively little aperture blockage.

However, the splasher type of feed exhibits a large difference in E and H plane beamwidths, typically 105 and 165 degrees, respectively, at the −10 dB points. If the E plane is matched to the dish, the H plane will be severely over-illuminated. If the H plane is properly illuminated, the E plane will be under-illuminated. Consequently this feed system is not recommended except in applications where gain and spillover are not critical.

The EIA feed[6] overcomes the shortcomings of the splasher by placing another dipole in the H plane as shown in **fig. 4B**. This helps to equalize the beamwidths as shown in **fig. 2**. This type of feed is one of the most popular types used on 70 cm EME.

The rectangular waveguide or horn (**fig. 4C**) is quite common in professional applications. It can be designed to have nearly equal E and H plane beamwidths. However, it is usually large and somewhat difficult to design, especially on the lower frequency bands. It's also more difficult to design if circular polarization is required. Typical design information is available in reference 7.

Relatively easy to design and build,[8] the cylindrical horn is often assembled from coffee cans or lengths of large-diameter tubing. Like the splasher, it has unequal E and H patterns. However, if a "choke" collar is placed near the mouth of the cylinder as shown in **fig. 4E**, the external surface radiation

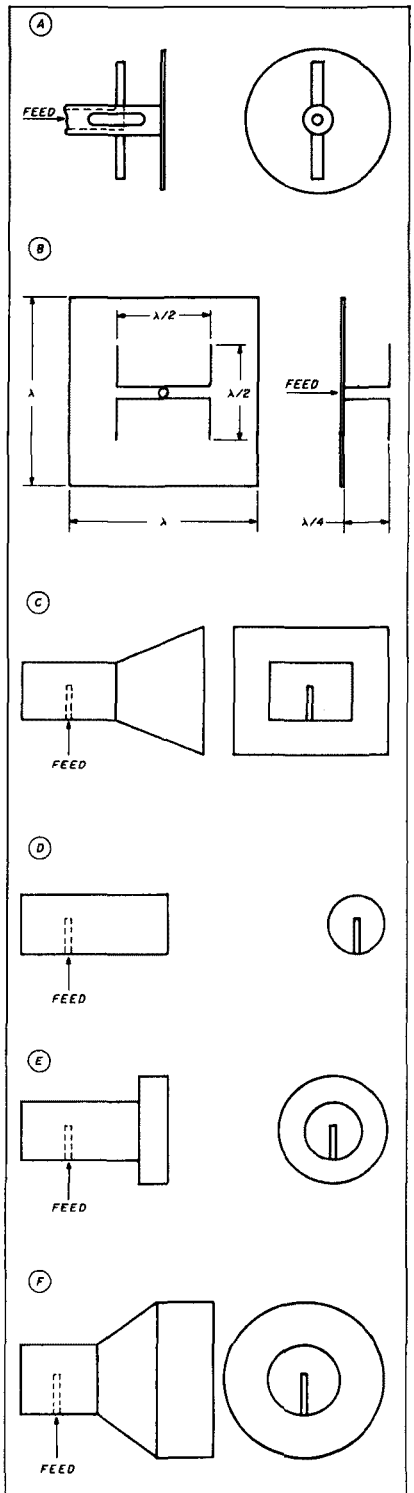


fig. 4. These are some of the representative feed systems used for parabolic antennas. *A*) splasher, *B*) EIA, *C*) rectangular waveguide/horn, *D*) circular waveguide/horn, *E*) circular waveguide/horn with flange, *F*) W2IMU dual-mode horn.

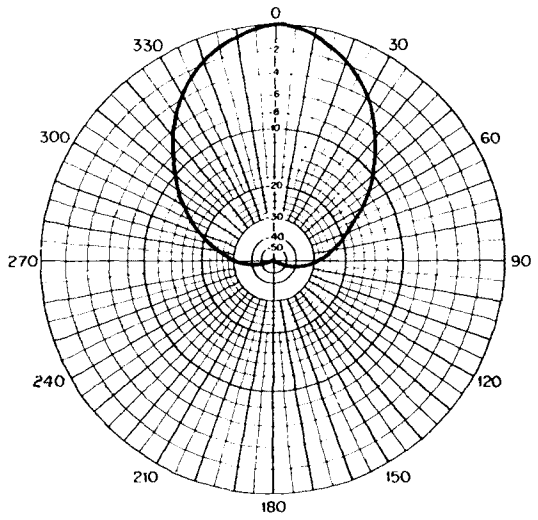


fig. 5. This is the approximate E plane pattern for the W2IMU dual-mode feed horn. The H plane beamwidth is virtually identical so it is not shown.

that causes this problem can be controlled. An improved circular waveguide feed with a choke is discussed in references 9, 10 & 11.

Dick Turrin, W2IMU, had patented a circular waveguide horn called the dual-mode feed.[12] Described in reference 3, it is shown in **fig. 4F**. The tapered section plus the larger diameter outer horn suppresses some of the other waveguide modes and forms symmetrical beamwidths as shown in **fig. 5**. This particular design is well suited for circular polarization on dishes with 0.5 to 0.6 f/d ratios and is presently the most widely used feed system on 23 cm EME.

Other feeds that are sometimes used by Amateurs include two-element quads or circular loops[12] and short Yagis.[13] K4QIF uses a quad loop in front of a plane reflector, which is well suited for a 0.45 f/d ratio parabolic antenna.

The least critical applications of parabolic antenna designs can use simple feeds such as a splasher quads, or the circular (coffee can) waveguide feeds. Where high performance is required, such as on EME, the EIA dual dipole, circular waveguide with chokes, or the dual mode circular polarization feed horns are preferred.

mechanical considerations

Numerous methods can be used for designing and constructing parabolic reflector dishes. The most popular techniques among Amateurs are the rigid solid dish, the stressed dish, and the trussed rib designs.

Solid surface and rigid dishes are very popular especially in commercial designs. Usually very accurate, they hold their shapes quite well. Many have perforated surfaces to allow some wind to pass through. However, the larger diameter units are usually very heavy — 500 to 2000 pounds (225 to 900 kg) — and can be quite expensive, even on the surplus market. One enterprising Amateur, WB6IOM, built his own version using honeycomb material for the reflector.[14] Recently the TVRO industry has manufactured dishes using petals that are often lighter in weight and less expensive than rigid designs. They are not always as strongly built as the others, but are usually sufficient for Amateur work through about 4.5 GHz.

At one time, stressed dishes were quite popular since they're relatively easy to build.[15,16] They can be tricky to handle and keep within tolerances when they exceed about 12 feet (3.7 meters) in diameter at 23 cm or 20 feet (6.1 meters) at 70 cm (432 MHz). To overcome these shortcomings, Amateurs have devised ingenious guy wire schemes that tension the surface and maintain the shape. However, this usually isn't sufficient, especially in extreme climates, so some Amateurs have even added 1 or more sets of guy lines behind the dish!

The preferred mechanical method of parabolic dish reflector design is the trussed rib type. There are many ways to accomplish this construction technique. Some use tubing, while others use angle stock. Basically this method is preferred, since it rigidly supports the rib and prevents distortion.

Some of the trussed methods are shown in **fig. 6**. **Figure 6A** shows the angle material method used by the engineers at NBS. The tubing method shown in **fig. 6B** was described in detail by VK3ATN.[17] The method shown in **fig. 6C** recently tried for TVRO applications, is quite easy to use for high f/d ratios and is particularly easy to build since only one curved surface is required.

The trussed rib design is recommended for Amateur applications.

Regardless of the construction method chosen, it should be strong. Additional rims should be used every 5 to 6 feet (2 meters) out from the dish center. Always use as many ribs as practical, since they increase stability and are likely to maintain tolerances more easily, especially on the reflector material. Twelve to 16 ribs should be a minimum, with 20 or more on the larger diameter dishes (i.e., those measuring over 20 feet or 6 meters). They should be attached to a large strong hub structure at the center of the dish.

feed mounting considerations

As mentioned, the feed size and weight are important. This is especially true on larger diameter dishes with a high f/d ratio, since the feed will have to be mounted very far from the reflector. This again emphasizes the need for a moderate size feed system that's as light as possible.

On small dishes with low f/d ratios, the feed can be mounted on a pipe or an arm with a "goose necked" shape. This is often found on commercial installations where the feed is integrated with a waveguide that also supports the feed system. In either case, some support or guying structure should be provided to hold the feed relatively close to its designated resting place and keep it firmly in place.

Probably the most common Amateur construction technique uses a tripod or quadpod mount to hold the feed system. In either case, adequate supporting structure is required behind the reflector to hold these supporting poles firmly in place. Failure to do so will incur unacceptable droop in the feed system. EMEers may also want to mount relays and a preamplifier behind the feed, which is an additional stress on the feed holding structure.

The feed mounting is an area that requires much forethought before the reflector is constructed. The distance from the feed to the reflector should be able to be moved in or out as well as up and down slightly for proper alignment, as will be discussed shortly. Don't forget that the support mechanisms also reduce the overall antenna gain by blocking the reflected signal.

phase center

One thing I didn't mention previously was the importance and use of the focal point. The calculated focal point of a parabolic antenna is where the phase center of the feed should be located. The phase center of a feed is the point from which the signal appears to emanate. Although it's not easily determined, it can be "guesstimated" with sufficient accuracy.

The phase center of the dual-dipole feed or a Yagi antenna is located near the front of the structure as shown in **figs. 7A and 7B**. The phase center of a horn is usually located a few inches/centimeters just inside the open end of a horn, and can be estimated using simple geometry as shown on **fig. 7C**, assuming that the angle shown is the same as the subtended angle of the dish being fed (see **ref. 1, fig. 6**).

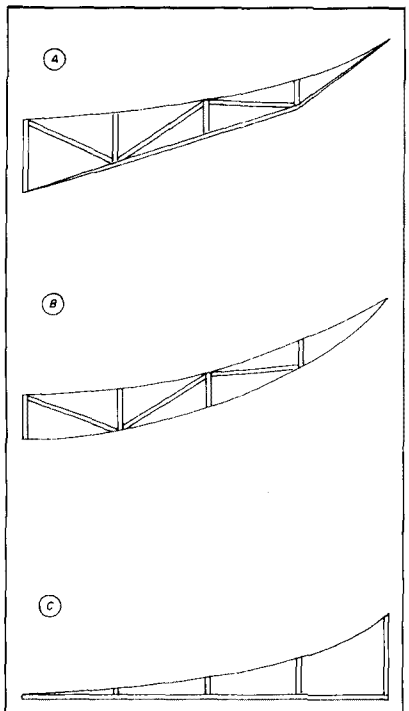


fig. 6. These are some typical methods of rib construction as described in the text and references: A) NBS with trussed ribs and straight stiffener. B) trussed rib per VK3ATN using two curved tubes. C) TVRO trussed rib using a single flat plane.

One feed system that I didn't mention previously is the log periodic array (LPA). It would appear at first glance to be a super feed, especially if wide bandwidth or several Amateur bands are to be covered.

However, in the LPA, the phase center is a function of frequency. Therefore, every time you vary frequency up or down, the phase center moves inversely. As a result, an LPA feed would be optimally matched to the reflector at one frequency only, a compromise that most Amateurs are unwilling to make. Furthermore, the E and H plane beamwidths of most LPA designs are usually quite dissimilar and therefore are not optimum, as discussed earlier.

Finally, the placement of the feed at the focal point is of only secondary importance. If the feed is located either too close or too far from the reflector, the gain may decrease slightly. If the feed is misaligned or skewed off axis, the main beam of the dish will shift slightly. This is normally not of great importance as long as the dish is moved an appropriate number of degrees one way or the other to compensate for any misalignment.

reflector design

So far the reflector has been mentioned only in passing. It's very important because even if the structure were sufficiently strong and mechanically perfect, a rough or distorted reflector surface could produce less gain.

As mentioned earlier, if any form of mesh or screen such as hardware cloth or chicken wire is used for the reflector, the hole size must be considered according to the frequency of operation, as previously discussed. One-inch (2.54 cm) chicken wire is often used through 23 cm. You can always upgrade the reflector later by replacing the screen with a mesh with smaller holes or by simply laying new mesh over the older mesh. This technique can be extended to just covering the inner portion of the reflector where most of the signal is concentrated.

The conductivity of the reflector surface is not overly important because the mesh is operating in a relatively high impedance (377-ohm) environment. Galvanized mesh is recommended. Rust won't necessarily affect performance unless the porosity of the surface begins to deteriorate. In fact, some experts say that a rusty mesh will have to virtually fall apart before it stops acting as a good reflector!

The rigidity of a mesh reflector must be such that it won't sag. This is partially a function of the number of ribs and supports. If there are too few ribs or supports, the mesh will have less support and hence be more likely to vary from the true parabolic shape.

A few other pointers are in order.

The mesh used for the reflector is normally purchased in rolls and must be formed to the parabolic shape. Depending on the size and placement of the mesh, it may be better to cut it up into smaller pie-shaped pieces and overlap each section, preferably at the ribs. The mesh should be tied off at numerous points on the ribs. Overlap is recommended. It should, if possible, be up to one quarter wavelength at the lowest frequency of operation[3]. Ohmic contact is not required.

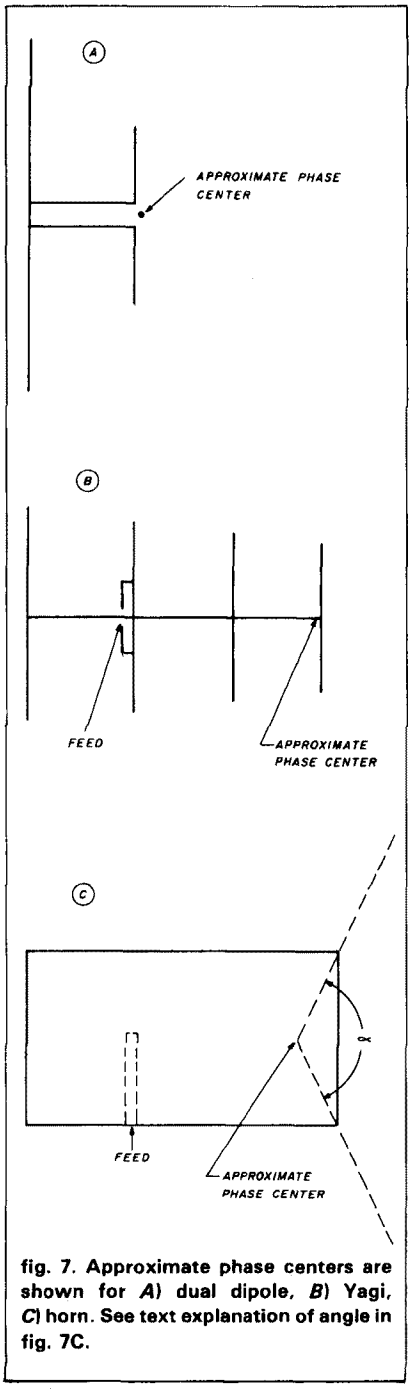


fig. 7. Approximate phase centers are shown for *A*) dual dipole, *B*) Yagi, *C*) horn. See text explanation of angle in fig. 7C.

Some individuals have attempted to improve performance by extending the reflector without changing the feed system. This can be quite counterproductive, especially if the reflector geometry doesn't exactly follow the parabolic curve. If the reflector is extended, the feed pattern must be widened accordingly. This is sometimes done when a person decides at a later date to extend a dish that is too small — but it can only be done if the feed is redesigned, too.

Dick Turrin, W2IMU, has suggested that a different tactic is more applicable.[18] If you're tempted to extend the reflector without changing the feed, then the reflector extension should be bent away at a right angle to the present reflector. In effect, this will deflect extraneous signals away instead of possibly bringing them in with improper phasing, thereby introducing noise pick or decreasing performance.

designing a parabolic dish antenna

We finally have most of the parameters in place, so let's get on with the design. First select the diameter of the reflector to be built. In order to do this, you must first determine the desired gain. Then consult the gain equations and/or **fig. 2** in reference 1 to determine the minimum diameter of the reflector.

Always choose a reflector diameter that will guarantee the desired gain. Err slightly on the larger diameter side to be sure your gain objectives will be met, but don't go overboard or the antenna will cover your entire backyard! Remember that every time you double the reflector diameter, the mechanical problems will at least quadruple!

Also remember that even if you can build the dish, you still have to place it on a mast and tower so you can rotate it! Furthermore, the reflector and feed mounting must be sufficiently rigid so that none of the mechanical dimensions change, either from variations in weather (temperature, wind, ice, etc.) or when the position of the antenna is rotated in either the azimuthal or elevation plane!

Before you can design the reflector, you must first decide which feed you will be using. Let's assume that you choose to use the EIA dual-dipole feed system. It will be necessary to determine the optimum f/d ratio for this feed.

Let's start by deciding that an f/d ratio of 0.5 is desired, as previously suggested. **Figure 6** in reference 1 shows that the subtended angle for a dish with an f/d ratio of 0.5 is approximately 106 degrees. Next determine the "space attenuation" at the rim of the reflector by consulting **fig. 7** in reference 1. We see that the edge taper for 53 degrees (one-half the subtended angle) is approximately 2 dB.

Next examine the pattern of the EIA feed in **fig. 2** of this month's column. At a beamwidth of ±53 degrees, the E plane beamwidth is down approximately 10 dB. This is not a great match for an f/d ratio of 0.5 since 10 dB plus the 2 dB edge taper adds up to 12 dB edge illumination, not the recommended 10 dB per reference 1.

An iterative procedure can be used if the 10 dB edge taper is not attained on the first try. At an f/d ratio of 0.55, the edge taper is approximately 1.8 dB at 50 degrees. At 50 degrees off boresight, the EIA pattern is approximately 8 dB below the main beam for a 10 dB total. This is a better choice for this feed system. However, any f/d ratio between 0.5 and 0.55 would probably be satisfactory for the EIA-type feed system.

You may have noticed that the H plane beamwidth of the EIA feed is narrower than the E plane, but this not be a significant problem, as discussed in reference 1. The overall gain will only be slightly less than the desired value because of under-illumination, but this is better than over-illumination!

For those interested in other feed systems, the W2IMU dual-mode horn is near optimum for an f/d ratio of 0.6.

You can determine this as we just did for the EIA feed by examining the pattern on **fig. 5**. Most of the circular horns in reference 9 are applicable for low (less than 0.5) f/d ratios.

evaluation of performance

The antenna evaluation methods mentioned in reference 6 are quite applicable to evaluating the performance of a parabolic dish type of antenna. The primary things to check are the beamwidth and the level of the first sidelobes. They should be close to the values shown on **fig. 3** and **table 2** in reference 1.

One other parameter, quite predominant in this type of antenna, is the possibility of skewing of the pattern off boresight. Most other types of antennas such as Yagis are quite straightforward in performance. All that's necessary is to look up at the boom and aim.

However, as just mentioned, a shift of the feed system of a reflector type of antenna may cause the main beam to be skewed slightly. Therefore, when evaluating a dish, first peak the signal and check that it's in the direction indicated on the rotator. If not, adjust the positioning of the feed or the rotator accordingly. For EME this can best be done by pointing the antenna at the Sun and peaking for maximum Sun noise.

improvements in performance

Time and space do not permit lengthy discussion of changes and improvements, but I will mention some briefly. Radomes are possible, but they can cause extra wind load and other problems. Most Amateurs simply stow their antennas during severe weather rather than risk storm damage.

Other feed systems than those described in this column will undoubtedly work, as long as they match the requirements outlined above. Likewise, new mechanical structures may be devised. Other variations may also be applicable if they meet the criteria presented here — for example, the use of an offset feed system.[18]

summary

The intent of this two-part series was to better acquaint the reader with the various aspects of designing a reflector type of antenna. Limitations of time and space, however, allowed us only to scrape the surface of parabolic dish antenna design.

As I stated earlier, the field of reflector antenna design is constantly being developed. Material is frequently being published on the subject and improvements are still being made. The greatest areas of improvement are in efficiency and low noise pickup. These improvements frequently require special feed designs, secondary reflectors or offset feed systems, considerations which are presently out of the reach of the typical Amateur. We can only hope that some day many of these improvements will be more easily adapted by Amateurs.

The material presented in these two articles should be more than sufficient for use in the design and construction of a good parabolic dish antenna. Build a strong structure, but keep weight down whenever possible. Use more ribs and inner rims than may initially appear to be necessary. Study the references cited. The rest is up to you. Good luck!

acknowledgements

I would particularly like to thank Dick Turrin, W2IMU, who over the last decade or so has shared his expertise and insight with me and so many others and thereby improved parabolic antenna design. I'd also like to thank Ed O'Connor, W2TTM, who prepared a list of questions he'd like answered in this two-part series. I've tried to cover as many of his suggestions as time and space allowed; any further ideas readers may have would be appreciated for use in future columns.

important VHF/UHF events:

March 1:	*EME perigee*
March 21:	*TE propagation optimum ±2 weeks*
March 28:	*EME perigee*
April 14:	*ARRL 144-MHz Sprint contest*
April 22:	*ARRL 220-MHz Sprint contest*
April 25-27:	*Dayton Hamvention*
April 30:	*ARRL 432-MHz Sprint contest*

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converting mobile microphones for handheld VHF transceivers

Many Amateurs purchase hand-held transceivers and use them for both portable and mobile operation. By adding additional external antennas, power amplifiers, speakers, and such, performance comparable to larger mobile rigs can be approached. Handheld units are considerably larger than mobile microphones, however, and can be inconvenient to use while driving. This is especially true when antenna cords, power cords, and speaker cords all run from different connectors on the handheld unit. Realizing this, manufacturers make special accessory microphones available; this article shows how to adapt readily available microphones — perhaps those already on hand — for mobile use at a considerable savings.

Most VHF rigs, including the popular ICOM 2A, have a single jack for an external microphone. When a suitable DC current is drawn from this connection, the transmitter is activated and an audio signal is applied to this same line. Thus, a microphone with a bias current is used to both turn on the transmitter and deliver audio.

The bias current is usually less than a milliamp and the available voltage will be about 3 volts (see **fig. 1**). The input impedance is then on the order of a few thousand ohms. **Table 1** outlines the measured characteristics of an ICOM 2AT.

By Edward A. Richley, KD8KZ, Bldg. 7, Apt. 201, 41 High Point Circle South, Naples, Florida 33940, and **Frank M. Caimi, WB3JCC,** P.O. Box 650163, Vero Beach, Florida 32965

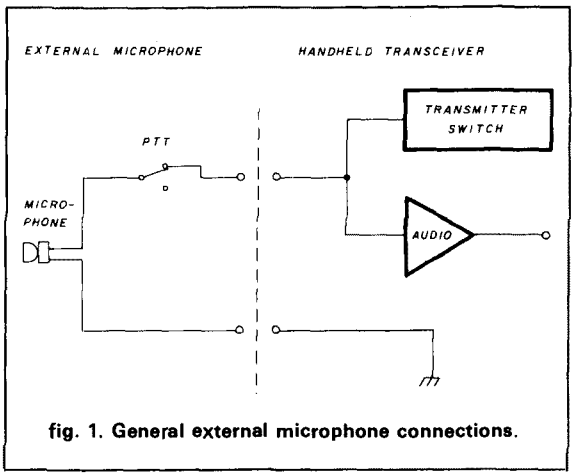


fig. 1. General external microphone connections.

table 1. Terminal characteristics of IC-2AT microphone input.

voltage	current
5.5	0*
4.1	74 μA
4.0	103 μA
3.5	270 μA

*Unit does not transmit.

Many high quality mobile microphones are available on the surplus market and at hamfests these days. Because they're often intended for a particular piece of commercial radio equipment, they may have electrical requirements that differ significantly from those outlined above. For example, flea markets have seen a flood of the Shure Brothers Model EM33G11 with a General Electric Label. This is a very comfortable, solid little microphone, complete with a Velcro® pad on the back (see **fig. 2**). Unfortunately, it's a dynamic microphone with very low impedances and a single switch contact. A variety of ceramic microphones, with extremely high output impedance, are also available for next to nothing (see **fig. 3**). We'll show how to modify both types at minimal cost — and in a single evening.

dynamic microphone modifications

A dynamic microphone is basically a low-impedance speaker operated as a microphone. The voltage output from such a microphone is only a millivolt or so. This must be raised to 100 mV into a load of a few thousand ohms for an adequate audio level. For this purpose, a small audio transformer could be used; this approach will work, but such transformers are quite



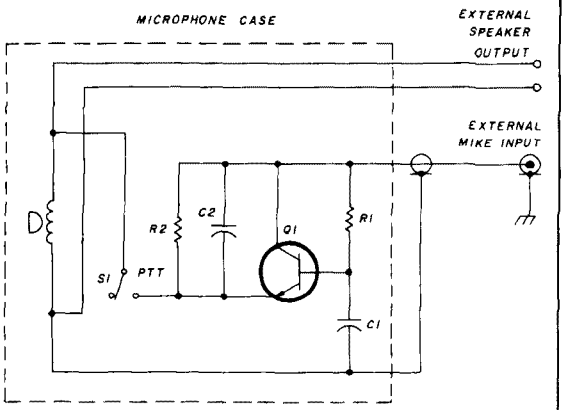
fig. 2. Dynamic microphone with modifications.



fig. 3. Ceramic microphone with modifications.

rare and don't allow for additional gain. For this reason, a solid-state approach has been selected and used with excellent results.

With a dynamic microphone, it's also possible to include a speaker/microphone in the same unit. The Shure EM33G11, for example, has only a single, normally open switch contact. This is sufficient for speaker/mike operation using the circuit shown in **fig. 4**. In this circuit, Q1 acts as a grounded-base amplifier with very low input impedance and adequate gain. Switch S1 is the push-to-talk button and serves to connect the amplifier as a load to the microphone input. When S1 is open, the dynamic element is connected as a speaker. When S1 is closed, enough bias current flows through Q1 to turn on the transmitter. C1 serves as a bypass to effectively ground the base to audio signals. C2 serves to ensure that radiated power does not cut off Q1. R2 affects the microphone gain and may need to be adjusted. The selection of Q1 should be made so that sufficient gain is available with only 1 mA of collector current. This choice is somewhat critical because the bias current is below 1 mA. Type 2N930 has a minimum h_{fe} of 150 at 1 mA and is commonly available. The 820 kilohm bias resistor creates about 550 μA of bias current to turn on the transmitter. Slight adjustment of this value may be necessary with different devices. h_{ib} is the input impedance in a common base configuration and is specified for the 2N930 as 25-32 ohms. This provides a slight mismatch to the 8-ohm dynamic microphone, but the gain of the device more than makes up for the difference. Furthermore, the output impedance of this



C1 1 μF tantalum
C2 0.001 μF
Q1 2N930 or similar device
R1 820 kilohm
R2 3 kilohm (see text)

fig. 4. Electronics for dynamic speaker/microphone.

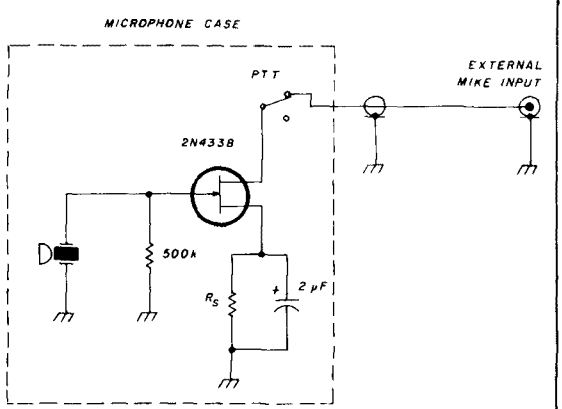


fig. 5. Ceramic microphone amplifier using a single n-Channel JFET.

device is at least 1 Megohm ($h_{ob} < 1$ μmho), so that an expression for the voltage gain is:

$$A_V = R_L/(R_{mic} + h_{ib}) \qquad (1)$$

With no additional load beyond that of the radio itself, the amplifier voltage gain is about 300. Thus, for a 1 mV signal from the microphone, a very substantial 300 mV will be available at the microphone input terminals. This was found to be more than adequate gain, and a further load of 3 kilohms was added in the form of R2.

The entire circuit of **fig. 4** can be built into the microphone body. The area between the speaker magnet and the shielding ring is more than adequate. Be sure to run the shielded lead of the cord from the amplifier output; this helps to reduce RF feedback. To make the dual phone plug connector at the radio, be sure to first make all cable connections to the plugs. (This is very important because once the two plugs are glued in place, there is no way to re-wire them). Then insert both plugs in their appropriate jacks and carefully fill the space between with epoxy to provide proper positioning. Let the epoxy set overnight. *Be careful to not glue the plugs to the radio*! When the glue has hardened, both plugs can be removed together as one dual unit.

ceramic microphone modification

Ceramic microphones operate on a completely different principle. In operation, sound pressure waves cause minute movement of a diaphragm which is mechanically coupled to a piezoelectric material. Induced mechanical stresses result in accumulation or depletion of charge between electrodes deposited on the material. Because the associated capacitance is relatively small, substantial voltages (typically as high as 100 mV) can be produced in proportion to the sound pressure input. Unlike the dynamic microphone, very little current is available for driving the load impedance since charge equalization between the electrodes requires relatively few electrons. Ceramic microphones are therefore very high impedance devices.

The design of an amplifier required to drive the transmitter input must take into consideration the microphone impedance, the load impedance, and whatever voltage gain is required to produce sufficient audio level. The ceramic microphone we used was a Johnson CB microphone purchased at a flea market for less than $3.00. Using an RMS-reading voltmeter, we determined an average output level of 35 mV into a 300 kilohm load. Since 250 mV RMS is the desirable input voltage to the IC-2AT, the required voltage gain for the proposed amplifier is 250 mV/35 mV = 7 or 17 dB.

Typical single-stage high input impedance amplifiers utilizing FETs can achieve the required gain. Voltage gain for the circuit shown in **fig. 5** is given by $A_V = g_m R_L$ where g_m is the FET transconductance at operating current and R_L is the AC load resistance.

Several trade-offs result when designing for high gain using the fixed load resistance of the ICOM. The AC load resistance (R_{input} = 10 kilohm) is approximately equal to the series/parallel combination of R52, R51, R55, R56, R57, and the base-emitter characteristic of Q23 (as indicated in the ICOM schematic provided with the 2AT), while the DC resistance is given by the data of **table 1**. A voltage margin at the FET drain sufficient to accommodate the amplified output signal requires a quiescent drain voltage several hundred millivolts below the supply voltage. As shown in **table 1** the open circuit supply voltage is 5 volts;

moreover, the unit will not transmit without a minimum current being drawn. As a consequence, the quiescent drain voltage must be several hundred millivolts below approximately 4 volts. A suitable operating point was chosen as V_D = 3.5 V, I_D = 270 μA.

These requirements pose a very severe set of constraints for the FET. The main problem is that a FET with a very low cut-off V_{GS} must be chosen. Very few FETs have a value of I_{DSS} below 1 mA. This is the maximum drain current that can be drawn. Thus, in our application, the device will have to be operated near cut-off. By writing a simple loop equation for the FET circuit, the constraint on cut-off voltage becomes clear:

$$V_{GS} + V_{DS} + I_D \cdot R_{DC} = 5\ V$$

or

$$V_{GS} < 5V - I_D \cdot R_{DC} = 3.5\ V$$

The FET must have a V_{GS} cut-off of less than 3.5 volts. A 2N4338 (or equivalent) meets this specification albeit with a lower transconductance than FETs with higher V_{GS} cut-off. The available transconductance is lowered further by operation at the 270 μA bias current available at the microphone terminals. Specified gate-to-source voltages, V_{GS} between 0 to −0.25 volt, yield a transconductance of 800 microsiemens for the desired operating conditions. Thus, the available voltage gain is given by 800 microsiemens x 10 kilohm = 8. Surprisingly, this meets the minimum design requirement previously specified for gain and provides adequate margin for AC voltage variations at the drain. The source resistor value R_S is adjusted to provide the 270 μA bias current and depends upon the manufacturer's tolerance spread for the 2N4338. A value less than 1 kilohm is probably optimal. Some variation of circuit gain is expected over temperature extremes.

This design proved to be operational but lacked flexibility of gain control and was therefore abandoned. In addition, most electronic junk boxes typically do not contain a 2N4338, and since other FETs will not function adequately, an alternate, though more complex, design was developed.

preferred design

Because operational amplifiers commonly provide high gain, noise immunity, low distortion, and stable characteristics, they're often used to replace discrete device amplifiers. The circuit shown in **fig. 6** utilizes a Bipolar/MOS op-amp in place of the FET of **fig. 5.** The op-amp can operate from supply voltages as low as 2 volts with current drain of only 350 μA and can therefore be powered from the ICOM microphone input circuitry.

Compared to conventional op-amps, the basic principle underlying operation of this design is uncommon. For this design, the op-amp quiescent current is supplied by the IC-2A upon closure of the PTT switch. Sufficient current flows to enable the transmit circuitry, thereby changing the mode from receive to transmit. Audio signals from the microphone are amplified by the voltage ratio R_F/R_I and appear at the output terminal. Only the AC portion of this signal flows in the load resistor R_A. The AC load current must, in turn, appear superimposed on the DC supply current. It is this AC power supply current that flows in the ICOM microphone circuit and causes modulation of the transmitter. Because the op-amp's power supply rejection ratio and CMRR are large ($>10^6$), voltage variations induced at the power terminals by the microphone signal do not induce unwanted feedback or gain variations.

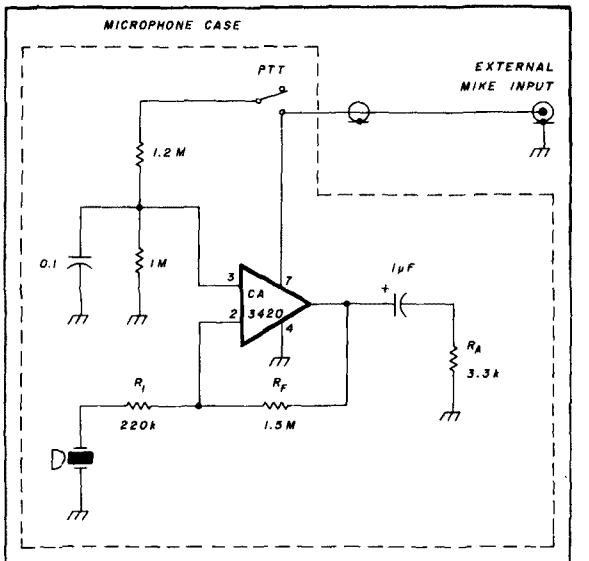


fig. 6. Versatile ceramic microphone amplifier using a low voltage, low power op-amp.

Typical quiescent current specifications of 350 μA for the CA3420 are sufficiently near the ideal DC supply current for the ICOM microphone circuit that the op-amp DC load need not be augmented. We measured the DC voltage at the power supply terminal and found ample headroom for the output signal.

As shown in **fig. 6**, operation within the common mode voltage range is guaranteed by the 1.2 and 1.0 Megohm bias resistors. This choice of operating point places the output stage in the center of its linear region. The 0.1 μF capacitor is necessary to remove the output signal from the input of the op-amp. The ratio of R_F to R_I, the input impedance of the radio, and the value of R_A determine the overall gain:

$$gain = (R_F/R_I)\ (R_{INPUT}/R_A) \qquad (2)$$

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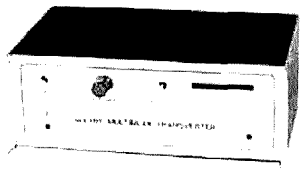
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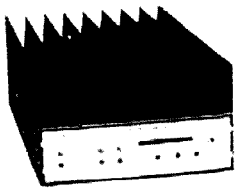
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R_A is typically 3.3 kilohm for satisfactory performance. Slight (plus or minus 6 dB) tailoring of the gain can be accomplished by adjusting R_A. In fact, less sensitive microphones can be used as long as the ratio of R_F to R_I is changed proportionately. Values of R_A below 1.5 kilohm are not recommended.

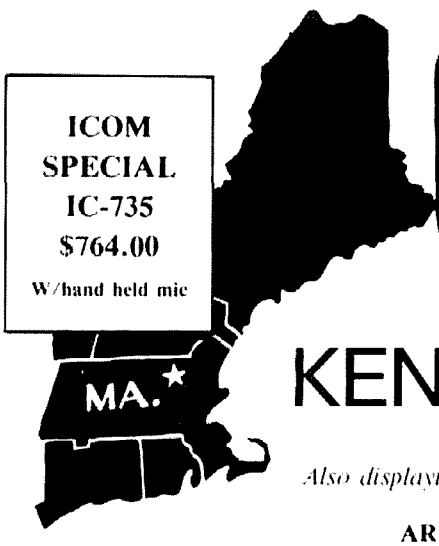
This circuit has been in use for over a year and has performed admirably in all types of weather.

summary

Although generally provided with electret microphones, both ceramic and dynamic microphones can be made to work with handheld transceivers by the addition of only a small amount of circuitry. In the case of dynamic microphones, a speaker/microphone combination can be easily constructed. Some variation in the circuitry may be necessary for different microphones and transceivers, but the general techniques should remain the same. These ideas should enable Amateurs to have a wide choice of affordable mobile microphones for use with existing transceivers.

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remote repeater programming
using a computer and a telephone

Upgrade your old controller with this simple Apple program

Back in 1983, the repeater association that I belong to decided that it was time to upgrade our 2-meter repeater controller. The old controller still was working well enough, but was really starting to show its age. The IDer used diodes to "memorize" the call sign. After the third change of call, the foils started to pull away from the PC board. We knew it was time for a change.

At the time, we really didn't have the money for a full-featured, controller, but a local ham, WA9FBO, was designing a controller that he planned to market. The controller was based on the Motorola 6809, a microprocessor with which he was familiar. It had all the necessities of a good repeater controller: autopatch, CW, voice ID, and reverse patch, to name a few. We decided to go ahead with WA9FBO's controller.

After the new controller was installed into a 1962-model G.E. Stationmaster, the fun began. The new controller has the capability of building "macros" for customizing the system to the users' tastes. To build these macros, you have to send standard DTMF tones to the controller, either over the air or telephone. Our group uses the telephone method, except during emergencies.

Every so often an improvement to the operating system required starting over again from scratch. Usually this meant no more than changing a couple of PROM chips. But doing this meant removing the power from the controller, thereby erasing all the macros. In our case, 45 minutes later, they were all reloaded and back in operation — if nobody made a mistake. We knew there had to be a better way!

While I suppose it would have been possible to tape record the tones for playback when needed, doing so would have required tape recording after each change to the system. To enter all the macros by hand presented an additional drawback because there aren't many telephones equipped with 16-button pads! What follows is what we feel is a much better solution.

a better way

In almost every Amateur Radio group, there are probably a few members with the required equipment to handle the task. What you'll need is someone with a computer that has autodialing capabilities. It can be any kind computer and any kind of dialer; mine is part of a modem that I use for information retrieval. Also, a nice thing to have, but by no means necessary, is a word processing program because it makes making changes a lot easier.

The program presented here (***fig. 1***) was written for Apple II series (except the IIc) computers with at least one disk drive and a Hayes Micromodem Iie (but not the MIcromodem II, which doesn't have touch-tone capabilities). All macro data is stored as standard Apple text files, such as those produced from word processors such as Apple Writer, Screen Writer and Apple Works (printed as a Text-ASCII file). It is written for DOS 3.3, but should be easily changed for PRODOS. The same basic idea of the program should be easy to convert to other computers.

When entering the program by hand, it isn't necessary to enter all the REM statements; they're included only for program clarity. If you're rewriting the program for another computer, remember to first get it to run, and then add the niceties such as screen formatting. And finally, if the computer doesn't support text files, or you don't have a word processor, DATA statements work just as well, but are not as easy to use in making changes.

As presented, the program is more or less a structured program and makes use of subroutines where allowed. The reason that the subroutines are at the front of the program is for program execution speed. The comments following the REM statements explain what the modules do.

The kind of performance that can be expected from the program depends on certain variables. Of greatest

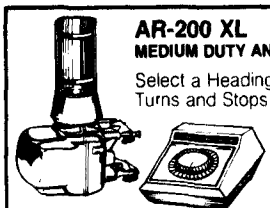
By John R. Kaiser, KD9BC, 803 Central Street, Oshkosh, Wisconsin 54901

```
1   REM    \-----------------------------------/
2   REM    !                                   !
3   REM    !  TOUCH TONE MACRO PROGRAM         !
4   REM    !  BY  JOHN R. KAISER   KD9BC       !
5   REM    !  COPYRIGHT 1985 BY J.R.K.         !
6   REM    !                                   !
7   REM    !-----------------------------------!
8   REM    !                                   !
9   REM    !  PROGRAM TO LOAD REPEATER MACROS  !
10  REM    !  OVER THE TELEPHONE LINES USING   !
11  REM    !  AN APPLE ][, ][+, OR ][e AND A   !
12  REM    !  HAYES MICROMODEM ][e             !
13  REM    !                                   !
14  REM    /-----------------------------------\
15  REM
16  REM    ***** GOTO START OF PROGRAM
17  REM
18  GOTO 77
19  REM
20  REM    ***** SUBROUTINE THAT FETCHES EACH NUMBER
21  REM
22  FOR F = 1 TO REC
23  LET S$ = S$(F): REM   GET SINGLE LINE FROM TOTAL FILE
24  FOR I = 1 TO  LEN (S$)
25  LET D1$ =  MID$ (S$,I,1): REM   GET SINGLE LETTER FROM LINE
26  REM
27  REM    ***** "PRINT" D1$ AT 3,33
28  POKE 1312,128 + ASC (D1$) : REM SCREEN MAPPED BYTE
29  GOSUB 36: REM  DIAL THE NUMBER
30  NEXT I
31  NEXT F
32  RETURN
33  REM
34  REM    ***** SUBROUTINE THAT DIALS THE NUMBER
35  REM
36  LET D =  - 1
37  FOR J = 1 TO 16: REM   CHECK FOR VALID NUMBER
38  IF D1$ =  MID$ (V$,J,1) THEN  LET D = J - 1
39  NEXT
40  IF D < 0 THEN  GOTO 47: REM  NOT A VALID NUMBER
41  POKE  - 16253 + SLOT * 16,D + 48: REM   DIAL NUMBER
42  FOR J = 1 TO DL
43  NEXT J
44  POKE  - 16253 + SLOT * 16,D + 32: REM   SILENCE BETWEEN NUMBERS
45  REM  FORJ=1 TO DL / 2: REM USE THESE TWO LINES
46  REM  NEXT J           : ONLY IF EXTRA DELAY IS NEEDED
47  RETURN
48  REM
49  REM    ***** READ THE TEXT FILE INTO S$( )
50  REM
51  LET J = 1
52  ONERR  GOTO 59
53  PRINT D$;"VERIFY";FILE$
54  PRINT D$;"OPEN ";FILE$
55  PRINT D$;"READ ";FILE$
56  INPUT S$(J)
57  LET J = J + 1
58  GOTO 55
59  PRINT D$;"CLOSE ";FILE$
60  POKE 216,0
61  IF  PEEK (222) = 6 THEN  GOTO 67: REM    FILE NOT FOUND
62  LET REC = J - 1: REM   NUMBER OF LINES IN FILE
63  GOTO 100: REM  GOTO MAIN PROGRAM
64  REM
65  REM    ***** FILE NOT FOUND ERROR MESSAGE
66  REM
67  PRINT "FILE NOT FOUND!"
68  PRINT "INSERT PROPER DISK AND PRESS RETURN.";
69  INPUT A$
70  GOTO 87: REM    GET NAME OF FILE
71  REM
72  REM    *****  START OF THE PROGRAM
73  REM
74  REM    *****  INITIALIZE VARIABLES
75  REM
76  LET DL = 30: REM   USED TO CONTROL SPEED OF DIAL, LESS = FASTER
77  LET SLOT = 2:  REM  THE SLOT NUMBER THAT THE MODEM IS IN
78  LET D$ =  CHR$ (4): REM   CONTROL-D USED FOR DISK ACCESS ON APPLE
79  LET D$ =  CHR$ (17): REM   CONTROL-Q USED FOR MODEM ACCESS ON HAYES
80  LET V$ = "D1234567890*#ABC": REM   USED FOR VALID CHECK
81  DIM S$(500): REM   500 LINES IN FILE-SIZE ACCORDINGLY
82  REM
83  REM    *****  GET THE NAME OF THE TEXT FILE
84  REM
85  TEXT
86  HOME
87  PRINT "ENTER THE NAME OF FILE TO WORK FROM"
88  INPUT FILE$
89  PRINT
90  PRINT "PLACE DISK WITH ";
91  INVERSE
92  PRINT FILE$
93  NORMAL
94  PRINT "IN DRIVE #1 AND PRESS RETURN.";
95  INPUT A$
96  GOTO 51: REM      READ IN TEXT FILE
97  REM
98  REM    ***** SET UP DISPLAY FOR THE REST OF THE PROGRAM
99  REM
100 HOME
101 PRINT "         S.O.L.A.R. MACRO LOADER"
102 PRINT
103 PRINT "       NUMBER BEING DIALED IS -->"
104 VTAB 5
105 HTAB 1
106 PRINT "----------------------------------------"
107 VTAB 7
108 POKE 34,6
109 REM
110 REM    ***** SET UP HAYES MICROMODEM ][e
111 REM
112 PRINT D$;"PR#2"
113 POKE  - 16253 + SLOT * 16,32
114 PRINT Q$;
115 REM
116 REM    ***** DIAL THE REPEATER
117 REM
118 LET S$ = "000-0000"
119 FOR I = 1 TO  LEN (S$)
120 LET D1$ =  MID$ (S$,I,1)
121 GOSUB 36
122 NEXT I
123 REM
124 REM    ***** WAIT FOR REPEATER TO ANSWER THE PHONE
125 REM
126 POKE  - 16253 + SLOT * 16,0
127 FOR J = 1 TO 4000
128 NEXT J
129 REM
130 REM     ***** SEND THE MACRO FILE
131 REM
132 GOSUB 22
133 REM
134 REM    ***** HANG UP THE PHONE
135 REM
136 PRINT
137 POKE  - 16253 + SLOT * 16,0
138 PRINT D$;"PR#2"
139 REM
140 REM    ***** THAT'S IT, RETURN SYSTEM TO NORMAL
141 REM
142 PRINT  CHR$ (8);"PR#0": REM   USE "PR#3" IF USING 80 COLUMNS
143 HOME
144 END
```

fig. 1. Touchtone program listing.

concern is the speed at which your local telephone company or the repeater controller will accept and process the tones. In the case of our group, the program is able to send a file of 3540 digits in 10 minutes and 45 seconds. That's with the variable DL set to 10 — not bad considering the relative slowness of Applesoft BASIC.

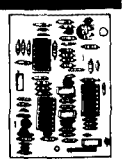
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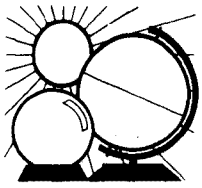
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How's the noise affecting your DX? The weakest DX signal that can be heard is limited by the noise level at your receiver. The readability is lowered considerably by that noise you hear. But where's the noise coming from?

One way of classifying noise sources is by identifying it as local or distant. Your receiver could be one local source of noise; yet most HF receivers used today, if kept in good repair, are good enough not to add appreciable noise that would affect amplification and detection of weak DX. Other sources of local electrical noise are power lines, electric motors, automobiles, televisions and computers, and solid-state type transmitters nearby.

If you're not located in the city near a factory, machine shop, or even in the country near a noisy power line, you're fortunate. If so, you've probably found, controlled, or scheduled around the worst of it. Finding local noise sources is an intriguing, frustrating form of transmitter hunting, an art on which several articles have been written.

If none of these local nuisances can be pinpointed as the source of your noise problem, then you'll have to consider the possibility of a more distant source — atmospheric noise. The cumulative effects of thunderstorms are propagated to your antenna via the same ionospheric layers you're using to communicate. The effects may also result from local storms within your ground wave propagation distance (50 to 100 miles or 80 to 160 km).

Potent thunderstorm noise at this time of year results from the following situation: In March and April, spring storms bring rain to much of the northern hemisphere. Fronts of warm and cold air generate the first major thunderstorms of the year. Fast-moving cold fronts produce particularly potent thunderstorms. As a storm front approaches your area, a significant increase in the noise level is heard. You'll first notice this increase at a one-hop distance away (about 600 to 1200 miles or 960 to 1920 km) when the storm front is about one day west of your location. The noise level will usually decrease after that until the storm reaches within a ground-wave's distance. Now, loud individual discharges can be heard. As the storm draws nearer, its sounds become part of the "local noise"; as it moves away, its noise decreases, then increases again as the front reaches the one-hop distance point a day or so later. (You can correlate this with storm progress reports on local television.) You can save time in looking for rare DX by tracking storms in order to pinpoint when the most favorable listening conditions are likely to occur.

last-minute forecast

The first and last weeks of the month are expected to produce the best higher frequency band, 15 to 30 meters, DX openings of the month. Whatever 27-day solar cycle variation is to occur, the rise in solar activity will take place during these days. Some transequatorial openings, mainly during geomagnetically disturbed periods can be expected. The middle two weeks can probably best be spent on the lower bands for DX openings to the east, north, and west during the night. During the disturbed magnetic periods keep your ear trained on the unusual rare DX you need for WAC. You may be lucky; March is the first month of the equinoxial geomagnetic disturbed period of each year, so you can expect disturbance and no real quiet in between. Expect also the intensity and length of the disturbances to increase throughout the month. April, however, is usually more disturbed than March. Spring equinox occurs on March 20th at 2203 UTC. The moon is full on the 3rd and at perigee on the 1st and 28th.

band-by-band summary

Ten and twelve meters, the highest day-only DX bands, are nearest the MUF for southern hemisphere paths. They will be open most days when the solar flux is above 75 during the 7 to 10-hour period centered on local noon. These bands open on paths toward the east and close toward the west. The paths are up to 2400 miles (4000 km) in single-hop length, and on occasion double that during evening transequatorial openings.

Fifteen meters, a day-only DX band open most of each day, has lower signal strengths and greater multipath variabilty than 10 and 12 meters. It will be best when the MUF is slightly above to slightly below this band — a transition period that occurs right after sunrise and just before sunset. Transequatorial openings will occur, with distances similar to 10 and 12 meters.

Twenty, thirty, and forty meters are both daytime and nighttime DX bands. Twenty is the maximum usable band for DX in the northern directions during the day. In combination with 30 meters, it provides nighttime paths for the day-only bands. Forty meters becomes the main over-the-pole DX daytime band, with some hours covered by 30 meters.

Eighty and one sixty meters, the night-only DX bands, exhibit short-skip propagation during daylight hours, then lengthen at dusk. These bands follow the darkness path, opening to the east just before local sunset, swinging more to the north-south near midnight, and ending up in the Pacific areas for a few hours before dawn. On some nights, 80 meters, with its higher signal strengths, will be the best band to use. One-sixty is also expected to provide good conditions most of the days. Please remember the DX windows of 3790 to 3800, 1825 to 1830, and 1850 to 1855 kHz.

MARCH

WESTERN USA

GMT	PST	N	NE	E	SE	S	SW	W	NW
0000	2:00	20	40	20	10	12	10	10	15
0100	3:00	20	40	20	10	12	10	10	15
0200	4:00	20	40	20	12	12	10	10	20
0300	5:00	20	40	20	12	12	10	10	20
0400	6:00	20	40	20	15	15	10	10	20
0500	7:00	30	40	20	20	20	12	12	30
0600	8:00	30	40	20	20	20	12	12	30
0700	9:00	20	40	20	12	12	10	10	20
0800	10:00	20	40	20	12	12	10	10	20
0900	11:00	40*	40	20	20	30	15	15	30
1000	12:00	40	40	20	20	30	20	20	40
1100	1:00	40	40	30	20	30	20	20	40
1200	2:00	40	40	30	20	30	20	20	40
1300	3:00	40	40	30	20	30	20	20	40
1400	4:00	40	30	20	20	30	20	20	40
1500	5:00	40	20	12	20	30	20	20	40
1600	6:00	30	20	10	15	40*	30	20	40
1700	7:00	30	20	10	12	20	30*	20	40
1800	8:00	40	20	10	10	20	30	20	40
1900	9:00	40	20	10	10	15	20	20	40
2000	10:00	40	20	10	10	15	20	20	40
2100	11:00	40	20	10	10	12	15	20	20
2200	12:00	40	20	12	10	12	12	15	20
2300	1:00	40	30	12	10	12	12	20	20
0000	2:00	40	30	15	10	12	10	12	20
0100	3:00	20	40	20	10	12	10	10	15

MID USA

	MST	N	NE	E	SE	S	SW	W	NW
	8:00	40	40	20	10	12	10	10	15
	9:00	30	40	20	12	12	10	10	20
	10:00	30	40	20	12	15*	10	10	20
	11:00	30	40	20	15	15	10	10	20
	12:00	40	40	20	20*	15	12	12	30
	1:00	40	40	20	20	30	20	20	40
	2:00	40	40	20	20	30	20	20	40
	3:00	40	30	20	20	30	20	20	40
	4:00	40	20	15	20	30	20	20	40
	5:00	20	20	12	20	30	20	20	40
	6:00	20	20	10	15	40*	30	20	40
	7:00	20	20	10	12	20	20	20	40
	8:00	20	20	10	10	20	20	20	40
	9:00	30	20	10	10	15	20	20	40
	10:00	40	20	10	10	15*	20	20	40
	11:00	40	20	10	10	12	20	20	40
	12:00	40	20	10	10	12	15	20	20
	1:00	40	30	12	10	12	12	15	20
	2:00	40	30	12	10	12	12	15	20
	3:00	40	30	15	10	12	10	12	20
	4:00	40	40	15	10	12	10	12	20
	5:00	40	40	15	10	12	10	12	20
	6:00	40	40	20	10	12	10	12	20

EASTERN USA

	CST	EST	N	NE	E	SE	S	SW	W	NW
	6:00	7:00	30	40	20	10	12	10	10	20
	7:00	8:00	30	40	20	12	12	10	10	20
	8:00	9:00	30	40	20	12	15	12	12	30
	9:00	10:00	40	40	20	15	15	15	15	30
	10:00	11:00	40	40	20	20	20	20*	20	30
	11:00	12:00	40	40	20	20	20	20	20	40
	12:00	1:00	40	40	20	20	30	20	20	40
	1:00	2:00	40	40	20	30	30	20	20	40
	2:00	3:00	40	40	20	30	30	20	20	40
	3:00	4:00	40	30	20	20	30	20	20	40
	4:00	5:00	40	30	20	15	20	20	20	40
	5:00	6:00	20	20	20	12	20	20	20	40
	6:00	7:00	20	20	20	10	20	20	20	40
	7:00	8:00	20	20	20	10	20	20	20	40
	8:00	9:00	20	20*	10	12*	20	20	20	40
	9:00	10:00	30	20	10	10	20*	30	20	40
	10:00	11:00	30	20	10	10	15	20	20	40
	11:00	12:00	40	20	10	10	15	20	20	40
	12:00	1:00	40	20	10	10	15	20	20	40
	1:00	2:00	40	20	10	10	12	20*	20	40
	2:00	3:00	40	20	10	10	12	15	20	30*

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.

*Look at next higher band for possible openings.

ham radio

product REVIEW

Butternut WARC Kit

The FCC recently opened the 24.890 MHz band (12 meters) for Amateur operation. Like many others, I began operation on the new band with a makeshift antenna and transmatch. In my case, the antenna was the HF6-V vertical manufactured by Butternut Electronics Corporation.

A few weeks later I obtained a Model A-18-24 WARC band conversion kit for the HF6-V. The kit consisted of two clamps and rod and capacitor assemblies which, when installed on the vertical radiator, would allow resonant operation on the 12- and 17-meter bands.

Each rod and capacitor assembly forms a parallel resonant circuit and decouples the length of the radiator above it in a manner similar to that of a conventional L-C trap. Installation of this kit has a negligible effect on the tuning of the antenna on the original six bands.

The kit required only about half an hour to install and "tweak." The SWR across the 12-meter band is less than 1.3:1.

Currently, we in the United States don't have permission to transmit on the 17-meter band; checks there have been limited to use of a noise bridge and my ears. The A-18-24 and HF6-V combination shows a definite resonance near 18.1 MHz. Signals from commercial stations in that region are also louder than with the unmodified antenna. I'm sure that readers in countries that have access to 17 meters can verify my observations with on-the-air QSO's.

—K1THP

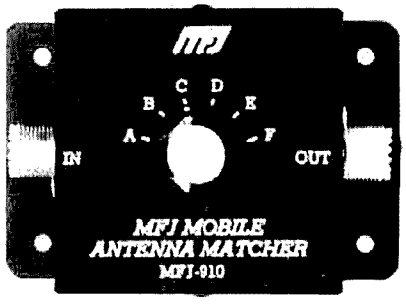
NEW products

mobile antenna matcher

MFJ Enterprises, Inc., has announced the introduction of the MFJ-910, a mobile antenna matcher. This device lowers SWR by capacitively matching a mobile antenna to 50 ohms, resulting in more power out of the transmitter and into the antenna, especially if a solid-state rig is being used. The MFJ-910 matches mobile antennas 10 through 80 meters.

One of the best features of this mobile antenna matcher is its compact size (only 2-1/2 × 2-1/2 inches) so it can be tucked into a corner of a car trunk or mounted inconspicuously with the mounting holes provided.

Installation is easy: simply connect the antenna to the SO-239 connector labeled "out" and the transmitter to the SO-239 connector labeled "in". This mobile antenna matcher retails for $19.95 each, plus $5 for shipping and handling.

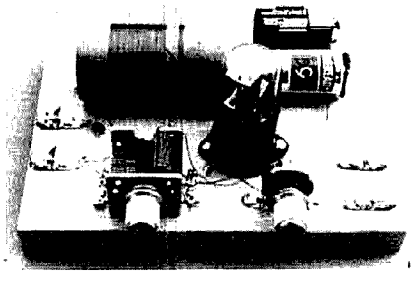


For more information, contact MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, Mississippi 39762.

Circle #318 on Reader Service Card.

one-tube radio kit

A new one-tube radio kit providing the experimenter or antique radio buff the opportunity to experience early radio construction and operation is available from Antique Electronic Supply.



The kit comes complete with breadboard, tube, and other parts. Batteries and headphones are optional. The kit and a 20-page catalog covering tubes, parts, books, and other items of interest to radio collectors and experimenters are available from George A. Fathauer, Antique Electronic Supply, 688 West First St., Tempe, Arizona 85281. The price of the kit is $16.95 plus $3.00 for shipping and handling.

custom fitted case for ICOM handhelds

Delta-Zulu has announced the release of its newest case, designed to fit the popular ICOM 2AT/3AT/4AT series of handheld transceivers and fill the need for a lightweight, compact case that allows user access to all controls. The new case, available in black or burgandy cloth-backed vinyl, features a unique zippered battery door that allows quick changing of standard sized battery packs and a transparent window for access to the touchtone pad. Its bottom seams are designed to improve stability when the unit is placed upright on a flat surface, reducing tipover.

Delta-Zulu cases come with a limited 90-day materials and workmanship warranty and are available from the manufacturer for $24.95 each. U.S. shipping is free.

For information, contact Crowley Manufacturing Co., 95 Federal St., Lynn, Massachusetts 01905.

Circle #317 on Reader Service Card.

meet the "gripmate"

Few jobs are more frustrating than those that require "three or more hands" — two to hold the work and a third or more to apply solder or perform some other operation. The smaller the component, the more difficult it usually proves to position it accurately and firmly.

The "GRIPMATE" is one of those why-didn't-I-think-of-it before? inventions — a clamp that provides up to *four* extra hands, each able to grip small components, wiring and other items in an infinite number of positions.



A base block clamps easily to a bench, cabinet or apparatus in any plane, carries four semi-rigid insulated wires (stayput arms), each fitted with an alligator clip to hold the job. Any of the arms can be replaced with one attached to a 48-mm diameter 2.5x safety magnifying glass for close-up work, or a small magnet when this is more appropriate than the clip.

The Super "GRIPMATE" kit, consisting of the

"HAM HOTLINE"
THE PROVEN MONEYMAKER

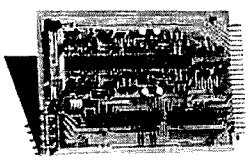
The **"Ham Hotline"** is a complete mailing list of novice amateur radio operators and current hams who have renewed, upgraded or modified their FCC licenses. These ham enthusiasts have proven to be excellent prospects for radio equipment, accessories and publications.

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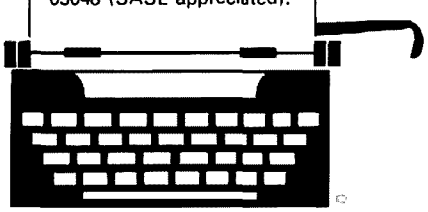
For more information call or write:

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Invitation to Authors

ham radio welcomes manuscripts from readers. If you have an idea for an article you'd like to have considered for publication, send for a free copy of the *ham radio* Author's Guide. Address your request to *ham radio*, Greenville, New Hampshire 03048 (SASE appreciated).



NEW products

base block, C-clamp, four 'stayput' arms, magnifier and magnet arms costs $19.95 plus sales tax (where applicable). Shipping and handling are included in the price.

For more information about "GRIPMATE" contact Gripmate Enterprises, Inc., P.O. Box 6179, Arlington, Virginia 22206-168.

Circle #316 on Reader Service Card.

sinadaptor

J.S. Technology, Inc. has announced the availability of a new product to expedite field or bench servicing of communications equipment. The sinadaptor model SAI-01 is a pocket-sized unit designed to plug directly into a VOM to provide sinad measuring capability anywhere radio systems require servicing.

The instrument is compatible with any VOM, AC VTVM, FET volt meter or other meter with AC measurement capability (analog or digital). The only requirements are that the meter have 2.5 VAC sensitivity or better and that the meter have a dB scale. The sinadaptor has been designed with very low output impedence so that the meter impedence does not effect accuracy.

By simply plugging the instrument into a VOM, connecting the VOM's test leads between the sinadaptor and the speaker leads of the receiver being tested, sinad measurement is displayed directly on the VOM's dB scale. For information, contact J.S. Technology, Inc., 39 Main Street, Scottsville, New York 14546.

Circle #315 on Reader Service Card.

new handheld

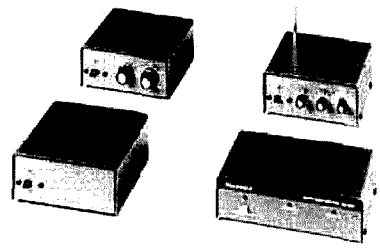
Encomm, Inc. has announced the newest addition to the SANTEC brand of radios. The ST-20T, described as a "smarter" handheld, features a large easy-to-read LCD atop a 32-function, 16-key keyboard. It provides two seven-digit auto-dial functions for automatically selecting and dialing through the repeater autopatch. All keyboard functions are accessible with one finger and minimal keystrokes, saving time and providing a very user-friendly interface. Its frequency range is 142-150.995 MHz for MARS and CAP users; other ranges from 140 to 176 MHz are available for export. Transmitter power is rated at 3.5 watts nominal and 5 watts maximum, depending upon the power supply voltage or battery used. Accessories and batteries are compatible with other popular models having similar slide-off battery packs.

For more information contact Encomm, Inc., 1506 Capital Ave, Plano, Texas 75074.

Circle #314 on Reader Service Card.

new from Heath

Four new Amateur Radio kits have been introduced by Heath Company. The HD-1420 Very Low Frequency (VLF) Converter allows a standard shortwave receiver to tune the 10 to 500 kHz band using the receiver's 3.5 to 4.0 MHz band. The HD-1422 Antenna Noise Bridge reveals the cause of any mismatch between a station's transmitter and its antenna. The HD-1424 Active SWL Antenna allows a shortwave radio to receive signals between 300 kHz to 30 MHz. And the HD-1530 Touch Tone Decoder is used in a series with the speaker of a receiver or scanner.



For a free copy of Heath's new catalog describing these new products and hundreds of others, contact Heath Company, Dept. 150-592, Benton Harbor, Michigan 49022. (In Canada write Heath Company, 1020 Islington Avenue, Dept. 3100, Toronto, Ontario, M8Z3.)

Circle #313 on Reader Service Card.

remote data interrogator for repeater tone panel

Communications Specialists has announced the availibility of its new DI-16 Data Interrogator. Designed for remote retrieval of accumulated time and hit information from the TP-38 Shared Repeater Tone Panel, it may also be used to remotely access the TP-38 for enabling or disabling repeater subscribers. The DI-16 interfaces with a simplex control base station on the repeater channel to establish a DTMF communications link with the TP-38 at the repeater site. Upon receipt of the proper command, the TP-38 will transpond data back to the DI-16 and show it on a four-digit LED display.



Provision is made for hard copy printout of the data on a Radio Shack Model TP-10 Serial Printer if desired. The DI-16 may also be connected to a host computer serial port for process-

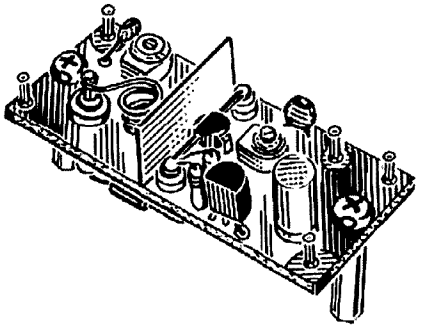
ing of the information for air time billing or other special purposes. All programmable functions of the TP-38 Shared Repeater Tone Panel may be performed remotely with the DI-16, eliminating the need for trips to the repeater site. The DI-16 uses a sixteen-digit keyboard and a high- visibility four-digit LED display. An automatic PTT keying circuit and an output port for a serial printer or computer adaptation are standard equipment. It is powered by 12 VDC and comes complete with 120 VAC power supply. Its size is 7.5 x 7.8 x 2.7 inches. The DI-16 is priced at $249.95 and is available for immediate shipment from factory stock. A full one-year warranty is included.

For information, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665-4296.

Circle #312 on Reader Service Card.

miniature GaAsFET preamp

Hamtronics, Inc., recently announced a new low-cost, miniature receiver preamp for the VHF and UHF bands. The model "LNW" is similar to the popular LNG GaAsFET preamp, which itself represented a breakthrough in GaAsFET preamp pricing.



The heart of the preamp is a new dual-gate GaAsFET that provides stable operation over a wide range of load impedances. Typical gain of the unit is 18 dB, and the typical noise figure is 0.8 dB. The unit measures only 5/8 x 1-5/8 inches. Mounting standoffs are provided to allow the preamp to be built into transceivers and receivers. It is easy to connect in series with the receive signal path using miniature coax.

Several models of the preamp are available:

model	nom frequency	tuning range
LNW-144	146 MHz	120-150 MHz
LNW-160	160 MHz	150-200 MHz
LNW-220	220 MHz	220-270 MHz
LNW-432	446 MHz	400-500 MHz

The price of the unit is only $19 in kit form and $34 assembled and tested.

A complete 40-page catalog including a full description of this and other products is available from Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535. (Include $1 for first-class mailing. For mailing overseas, enclose $2).

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THE JOY OF QRP: STRATEGY FOR SUCCESS

The experts reviewed Ade Weiss's (WØRSP) book and wrote: **George Dobbs, G3RJV**, *Radcom* "I am most impressed both with the scope and content of the book . . . the sections on objectives, planning, operating techniques, band selection and propagation would help any Amateur whatever power is being used — *SPRAT* "a comprehensive guide to the whole subject of QRP . . . a great book for QRP'rs and a lot of QRO operators would benefit from reading it." **Doug DeMaw, W1FB**, *QST* "I found the book easy to read, and the text is interesting throughout. I would have no hesitation in recommending WØRSP's book to any Amateur interested in QRP operation. In fact, it will provide great reading for nearly any active ham." **Bill Welsh, W6DDB**, Novice Ed., *CQ* "151 pages covering QRP from basics to fine points in 8 interesting chapters . . . Novices will have no difficulty understanding the explanations." **Fred Bonavita, W5QJM**, *QRP Quarterly* "In no other place have I encountered such a well-founded statement of the philosophy of QRP operating."

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MMT220/28	220ZMHZ,28 or 50 IF, 15W	**$219**
MMT144/28	144MHZ,28 or 50 IF,15W	**$179**
MMT144/28R	144MHZ,28IF,GAASFET,DBM,25W	**$299**
MMC144/28HP	144MHZ,28IF,GAASFET,DBM	**$ 79**

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MV1296	1296MHZ,GAASFET,1.0DB NF,SW	**$269**
DX432	432MHZ,GAASFET,0.5 DB NF	**$129**
MV432S01	432MHZ,GAASFET,0.7DB NF,SW	**$239**
DX220	220MHZ,GAASFET,0.4DB NF	**$129**
MV220S01	220MHZ,GAASFET,0.5DB NF, SW	**$229**
OX144	144MHZ,GAASFET,0.3DB NF	**$129**
MV144S01	144MHZ,GAASFET,0.5DB NF, SW	**$229**
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NEW products

software toolbox

Communcations Electronics Specialists Inc. (CES) has announced the release of the CES BeamCalc Software Toolbox for Commodore 64 and Apple II computers. The program assists Amateur Radio Operators with beam heading calculations from their location to locations in most areas of the world. The operator simply enters his latitude and longitude and then selects a specific location. The program provides the beam heading in degrees relative to true north, latitude and longitude of the selected location, distances in statute and nautical miles, time difference from GMT, DX Zone number, postal rates, numbers of IRC coupons, and QSL bureau information. Included in the CES BeamCalc Toolbox are program segments for radio line-of-sight calculations, satellite tracking, moon tracking, and solar outage predictions.

CES BeamCalc is menu-driven and comes with a detailed operating guide.

For information, contact CES, Inc., 803C South Orlando Ave., Winter Park, Florida 32789.

Circle #311 on Reader Service Card.

RF design programs

RF Notes No. 2 is the second in a series of RF design aid programs for problems frequently experienced in radio frequency design. RF Notes No. 2 contains four programs that can aid in such activities as inductor design, including design of single-layer coils (both close and space-wound) and toroid coil design. Complex impedance matching circuit design includes L, PI, T, and wideband configurations. In addition, a short program for capacitor applications evaluation is included. The price is $60.00. Designed for the color or monochrome IBM-PC with 128K, RF Notes No. 2 requires a graphics card.

For information, contact Etron RF Enterprises, P.O. Box 4042, Diamond Bar, California 91765.

Circle #310 on Reader Service Card.

DTMF encoder/decoder

Cetec Vega's new Model ED-707 mobile encoder/decoder is built to endure hard use. Designed for tough service, it's rugged and dependable, and includes a comprehensive selection of operational features including the following: selective, group, and all-call (2 to 6-digit messages); single-tone transpond; full DIP-switch programmability; command reset; audible encode sidetone; fully sealed keypad with 12 DTMF digits; horn switch, speaker switch, and new "T" switch; and wrong-digit reset or wrong-digit lockout (jumper selectable).

The fully sealed keypad has a positive tactile feel. It encodes 12 DTMF characters with audible sidetone beep, while producing a PTT output (with preset off-delay period). Also included are speaker, horn, reset, and "**T**" switches, with LEDs to indicate the state of the speaker and horn switches. Adjustable timers are provided for sounder/horn, tone-burst transpond, off delay, PTT delay, and interdigit time.

For details, contact Cetec Vega, 9900 Baldwin Place, El Monte, California 91731.

Circle #309 on Reader Service Card.

multicomm 2.0

MULTICOMM 2.0, an updated version of MULTICOMM 1.0 is now available from Multi-Comm Communications Software, Inc. MULTICOMM 2.0 includes Morse code transmit capability, a faster menu screen, user-definable function keys, and a terminal screen memory feature.

The new program includes ASCII, RTTY, and Morse Code Communications, with full teletype (Baudot) emulation, definable function keys, an Active Menu, file transmit and receive, and simultaneous printing, and many other features, for $49.95.

IBM-PC (and Compaq) system requirements are: 128K RAM, 1 disk drive and an 80 column monitor. RTTY communications requires a terminal unit with an RS-232 interface. Telephone communications requires a standard modem.

MULTICOMM 2.0 is available for a limited time at only $39.95.

For further details, contact Multicomm Communications Software, 1806 Foxwood Drive, Houston, Texas 77008.

Circle #308 on Reader Service Card.

circuit design/analysis software

Hayward Electronic Systems, Inc., has announced that it will supply LADPAC, a software package for the design and analysis of radio frequency ladder circuits. Ladder networks are used extensively for electronic filtering and impedance matching.

LADPAC will design lowpass, highpass, traditional bandpass, coupled resonator LC bandpass, and crystal ladder filters. A special feature is a routine for the design of pseudo-elliptic lowpass filters. LADPAC analyzes networks for transducer gain, impedance match (return loss), phase, and group delay. Impedances may be plotted on a built-in Smith chart. A special routine draws and edits the schematics of circuits designed by other LADPAC programs.

Hardware requirements for running LADPAC are an IBM-PC or compatible, DOS 2.0 (or later

version), at least one 5.25-inch disk drive, 192 kilobytes of RAM, and a color graphics adapter. A dot-matrix printer is recommended.

LADPAC and LADPAC 87, which provides enhanced operating speed (but requires the 8087 co-processor), are each priced at $149 each. For further details, contact Hayward Electronic Systems, 7700 SW Danielle Avenue, Beaverton, Oregon 97005.

Circle #306 on Reader Service Card.

logging software

The Log, the latest Amateur Radio software from Cynwyn, runs on the 16K or 32K Radio Shack Color Computer and is available in cassette or RD-DOS versions. When loaded into the computer, it tests for memory and configures itself to the maximum advantage of the available memory automatically. It provides fields for entering all standard logbook data as well as a "Remarks" field. The program — which will search log entries by call sign, QTH, and date — allows operators to obtain hard-copy printouts of their logbooks. The user with a 32K disk system can store up to 13,600 contacts per disk.

The log is available for $19.95 per cassette and $24.95 per disk. (Add $2.00 for shipping. Call signs should be included with orders.)

For information, contact Cynwyn, Suite 2F, 4791 Broadway, New York, New York 10034.

Circle #307 on Reader Service Card.

metroplex network

Metroplex, a key link in the North American Teleconference Net, is now on the air with its satellite broadcast news service, specially designed for transmission on Amateur repeaters and remote bases.

The Metroplex Network carries three exciting programs, co-produced by Metroplex and Westlink Rdio, for the Amateur Radio Community: an up-to-the-minute national news program and a national Amateur Radio swap-and-shop program every week, as well as all four North American Teleconference Radio Nets (NATRNs) each year.

Network-affiliated club stations downlink the programs from a commercial satellite transponder for a very low annual fee or receive the programs via a UHF link or a telephone company dialup line. Many clubs already have the basic antenna systems that are required.

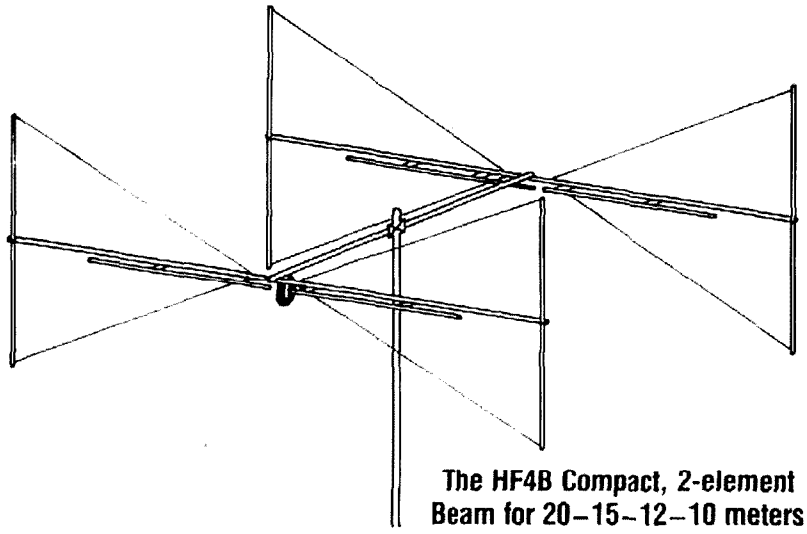
The total package is available every Monday evening at 10 PM EST/7 PM PST and Friday nights for NATRN.

All news and production personnel are working professionals who have volunteered their time for this informative way to help unite the world of Amateur Radio.

To become a network affiliate, or to learn more about it, contact Alex Magocsi, WB2MGB, at the Metroplex Network, Leonia, NJ 07605 or call 201-592-7614 for recorded information.

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The HF4B **BUTTERFLY** ™ has not sacrificed performance for compactness. Its unique design with tanned elements and L-C circuits avoids use of power-robbing traps yet provided high-efficiency operating on all bands. The **BUTTERFLY** ™ outperforms anything in its class.

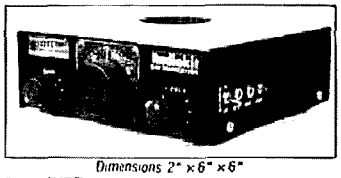
The HF4B offers an SWR of 1.5:1 or less at resonance. Its 2:1 bandwidth is 200 kHz on 20 meters, 450 kHz on 15, 1.7 MHz on 10, and across the entire 12 meter band. And it will handle the legal power limits both CW & SSB. Gain is at least 3 dB on 20, 4.5 dB on 15 and 5 dB on 10 & 12 meters. Front-to-back is up to 18 dB on 10, 12 and 20m, and up to 15 dB on 15m.

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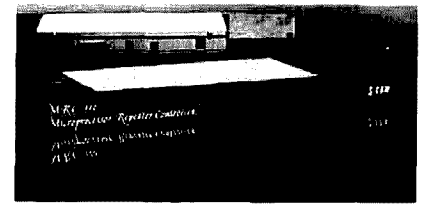
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The MRC-100 will support a second control receiver and comes with a power adapter for battery backup. A full two-year warranty is included.

For information contact S-COM Industries, P.O. Box 8921, Fort Collins, CO 80525.

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new catalog

The 1986 Grove Enterprises catalog of receiving accessories and publications is now available. Grove, a leading mail-order discount dealer of scanners and short-wave receivers, also manufactures preamplifiers, antennas, tuners, filters and other accessories for listening. Grove publishes books in the field and also *Monitoring Times*, a 40-page monthly tabloid covering all aspects of monitoring the radio spectrum.

For a free copy of the new Grove catalog and a sample of *Monitoring Times*, write Grove Enterprises, P.O. Box 98, Brasstown, North Carolina 28902.

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flea market

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SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

SCHEMATICS: Radio receivers 1920's/60's. Send brand name, model No., SASE. Scaramella, PO Box 1, Woonsocket, RI 02895-0001.

ANTIQUE RADIO CLASSIFIED "The National Publication for Buyers and Sellers of Old Radios and Related Items". Published monthly—features articles, regular columns, photos, and ads. Sample copy $1.50. ARC, 9511-23, Sunrise, Cleveland, Ohio 44133.

IBM-PC RTTY/CW CompRtty II is the complete RTTY/CW program for the IBM-PC and compatibles. Virtually any speed ASCII, BAUDOT, CW. Text entry via built-in editor. 10,000 character transmit/receive buffers. Adjustable split screen display. Instant mode/speed change. Hardcopy, diskcopy, break-in buffer, select calling, text file transfer, customizable full screen logging, 24 programmable 1000 character messages. Now with *WRU (simple mailbox). Ideal for MARS and traffic handling.* Requires 128k PC, XT, AT, PCjr, PC-DOS, serial port, RS-232CTU $65. Send call letters (including MARS) with order. Mention Ham Radio Magazine and take $10 off. David A. Rice, KC2HO, 7373 Jessica Drive, North Syracuse, NY 13212.

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COMING EVENTS

Activities — "Places to go . . ."

CALIFORNIA:FCC exams, Novice-Extra. Sunnyvale VEC ARC. (408) 255-9000 24 hour. 73, Gordon, W6NLG, VEC

OHIO: The all new 17th Annual B*A*S*H — New Location, new entertainment, new food — will be held on Friday night of the Hamvention, April 25, 1986. The new location is in the Conference Center at the HARA Arena and Exhibition Center, (the same location as the Hamvention) starting at 7 PM. There is no admission charge, and free continuous entertainment. Food and beverages available. Two exciting top awards, and many others. Stay right at HARA when the Hamvention closes on Friday evening and meet your friends and join us for an evening of fun and entertainment. Sponsored by the Miami Valley FM Association, POBox 263, Dayton, Ohio 45401.

MASSACHUSETTS: Chelsea Civil Defense will sponsor Amateur Radio evening classes at Chelsea High School starting March 18, 1986. Novice, basic level ham license, Technician or General license. Minimal charge for materials only. For more information about Amateur Radio and classes write: Frank *Masucci,* 136 *Grove Street,* Chelsea, MA 02150. Please include phone number.

INDIANA: The Morgan County Amateur Radio Association is sponsoring the Indiana Hamfest, March 9, Indiana State Fairgrounds Pavillion Building, Indianapolis. Admission $5.00 at the door. Flea market table $8.00 by reservation only. Flea market space, no table, $3.00. Reserved table setup Saturday, March 8, 3-9 PM. Space setup Sunday, March 9, 6-8 AM. Free parking. Talk in on 145.25. For table reservations SASE before March 1 *to Aileen Scales,* KC9YA, *3142 Market Place, Bloomington,* IN 47401 (812) 339 4446.

FLORIDA: 16th North Florida Ham/Swapfest, Saturday and Sunday, March 15-16, Fort Walton Beach. Admission $3.00 at door. XYLs and kids free. ARRL/VEC exams Saturday. Indoor flea market and dealers. RV parking. contact Hud Huddleston, 925 Forest Avenue, Fort Walton Beach, FL 32548. (904) 862-2566.

NEW JERSEY: *The Delaware Valley Radi Association's 14th* annual Flea Market, Amateur Radio and Computer equipment, Sunday, March 23, 8 AM to 2 PM, New JerseyNational Guard 112th Field ARtillery Armory, Eggerts Crossing Road, Lawrence Township. Advance registration $3.00 by March 10. $4.00 at door. Sellers bring own tables. Talk in on 146.07-.67 repeater. For registration and/or spaces SASE to KB2ZY, Box 441B, RD -1, Stockton, NJ 08559.

NEW JERSEY: The Old Bridge Amateur Radio Association's annual indoor/outdoor electronic Flea Market, *April 6,* Knights of Columbus Hall, Pine Street, Old Bridge. Sellers contact Bob Navin, N2DFJ, at (201) 251-6449 or Sam Ugliarolo, WB2HAE, at *(201) 583-2158. Talk in on 147.120 and 146.520.*

NEW YORK: The Westchester Emergency Communications Association's 2nd annual Hamfest and Electronics Fair, WECAFEST '86, Sunday, March 16, Westchester Community College, Valhalla. General admission $2.00. People under 16 admitted free. Flea market, ham radio equipment testing, FCC license exams and more. Plenty of free parking. Dealer booths $10 each by reservation only. For reservations or information: WECAFEST '86, PO Box 348, Millwood, NY 10546.

GEORGIA: The Columbus, GA, ARC will sponsor a Hamfest, Saturday, March 22, 9 AM to 5 PM and Sunday, March 23, 9 AM to 3:30 PM, Columbus Municipal Auditorium, Columbus. Flea market, free coffee and free parking for self-contained RVs (no hookups). Activities for spouses and kids. Indoor tables $6.50 per day. Contact Joe Mills at (404) 687-3050 or CARC, PO Box 6336, Columbus, GA 31906 for table rentals. FCC exams given Saturday morning 9 AM EST at the Colonial Inn. Contact Billy Bedgood (404) 323-5579 for exam information. Details on Saturday night buffet dinner available at Hamfest site. Talk in on 146.01/61. For more information: Billy Bedgood (404) 323-5579 or Bill Haycock (404) 327-7345 or CARC at address above.

NEW JERSEY: The Splitrock Amateur Radio Association's first annual evening Hamfest, March 14, VFW Post 3401, Tabor Road, Morris Plains. Starts 7 PM. Admission $1.00. Sellers tables $7.00. Suggest advance booking. Free parking. Talk in on 146.385/146.985. For information: SARA, PO Box 3, Whippany, NJ 07981.

WISCONSIN: The Madison Area Repeater Association (MARA) announces its 14th annual Swapfest, Sunday, April 6, Dane County Exposition Center Forum Building, Madison. Doors open 7:30 AM for flea market sellers and 8 AM for public. Admission $2.50 advance and $3.00 at the door. Children 12 and under admitted free. Flea market tables $5.00 each advance, $6.00 at the door plus admission. Advance tickets and tables by March 31, 1986. FCC exams by Four Lakes ARC during Swapfest. Preregistration requested but walkins will be accepted. For information/reservations: MARA, PO Box 3403, Madison, WI 53704 or (608) 222-4744.

KENTUCKY: The Lincoln Trail Amateur Radio Club's Hamfest, March 22, Prichard Community Center, Elizabethtown. Talk in on 146.38/98. For information or reservations: Hubert Hensley, WD4GDA, PO Box 387, Vine Grove, KY 40175 (502) 877*2234.

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NEW HAMPSHIRE: The annual Interstate Repeater Society's Flea Market, Saturday, March 15, Hudson Lions Club, Lions Avenue, Hudson. 9 AM to 4 PM. Doors open 8 AM. Admission $1.00. Tables $8.00 each. Free parking. Talk in on 25/85 and 52. For information or table reservations: Richard everhart, WB8YGR, 25 Briand Drive, Nashua, NH 03063. (603) 889-3479.

NEW JERSEY: Shore Points ARC invites everyone to Springfest '86, Saturday, March22, 9 AM to 2 PM, Atlantic County 4H Center, Egg Harbor City. Indoor selling space and covered tailgating. Sellers $5.00 per day (bring own table). Buyers $2.50 advance, $3.00 at door. Talk in on 146.985 and .52. For information: SPARC, PO Box 142, Absecon, NJ 08201.

MINNESOTA: Rochester area Hamfest, Saturday, April 5, John Adams Junior High School, 1525 NW 31st St, Rochester. Doors open 8:30 AM. Large indoor flea market, radio and electronics items. Refreshments and plenty of free parking. Talk-in on 146.22/.82. WOMXW. For information: RARC, WB0YEE, 2253 Nordic Ct. NW, Rochester, MN 55901.

ILLINOIS: Lamarsfest '86, sponsored by the Libertyville &Mundelein ARS. March 23, Lake County Fairgrounds, Grayslake. Doors open 8 AM. Large indoor electronic and radio swapfest. Tables available by reservation. Donation $2.00 advance, $3.00 door. Talk in on 146.52 simplex. Waukegan Rptr 147.63-03. For information or reservations: LAMARS, Box 751, Libertyville, IL 60048.

MARYLAND: The Baltimore Amateur Radio Club (BARC) will present the 1986 Greater Baltimore Hamboree and Computerfest, April 6, Maryland State Fairgrounds Exhibition Complex, Timonium. Indoor flea market, large dealers display area. Outdoor tailgating. Food service, free parking. Gates open 8 AM. Admission $4.00. Children under 12 free. For information: GBH&C, PO Box 95, Timonium, MD 21093|0095. 1301| 561-1282.

NEW JERSEY: The Chestnut Ridge Radio Club's 9th annual Flea Market, Saturday, March 22, Education Building, Saddle River Reformed Church, East Saddel River Road and Weiss Road, Upper Saddle River. Tables $10.00 first, $5.00 each additional. Tailgating $5.00. Contact: Jack Meagher, W2EHD. (201) 768-8360.

MICHIGAN: 25th annual Michigan Crossroads Hamfest sponsored by the Southern Michigan Amateur Radio Society, Saturday, March 22, Marshall High School, Marshall. 8 AM to 3 PM. Setup 6:30 AM. Tickets $2.00 at door, $1.50 advance. Table space 50 cents/foot. For reservations SASE to SMARS, PO Box 934, Battle Creek, MI 49016 or phone Wes Chaney, N8BDM (616) 979-3433. Amateur Radio license exams, Novice through Extra, given at Hamfest. Preregistration requested but walkins will be accepted. Send Form 610, SASE $4.00, check or MO to ARRL/VEC: License Exam, PO Box 2, Pleasant Lake, MI 49272.

MASSACHUSETTS: The MIT UHF Repeater Association and the MIt Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday, March 19, 1986,7 PM, MIT Room 1-134, 77Mass Ave, Cambridge, MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-5820/646-1641 or Craige Rodgers at 225-6616. Exam fee $4.00. Bring copy of current license, 2 forms of picture ID and completed form 610 (available from FCC in Boston: 223-6609).

MICHIGAN: The South Eastern Michigan Amateur Radio Association (SEMARA) will hold its 28th annual Hamfest Swap and Shop, April 6, 8 AM to 3 PM, Grosse Pointe North High School, 707 Vernier Road,Grosse Pointe Woods. Advance tickets $1.00. At the door $3.00. Advance tables $8.00. At the door $10.00. ARRL, DX and RTTY forums. Talk in on 147.75/.15 and 146.52. For information: SEMARA Hamfest, PO Box646, St. Clair Shores, MI 48080 or phone Fred Lewis NK8M (313) 881-0187.

OPERATING EVENTS

"Things to do . . ."

MARCH 2: From Alamo Plaza, San Antonio, Texas, next to the Alamo, W5SC will operate a special event station to celebrate Texas Independence Day as part of the Texas Sesquicentennial activities. 0600Z to 0600Z. 10, 15 and 20 meters, lower portions of General CW and voice bands. QSL and No. 10 SASE to W5SC, 90 Brees Blvd., San Antonio, TX 78209.

MARCH 22 The *Macon Amateur Radio Club*, W4BKM, will operate special event station from the Cherry Blossom Festival in Macon, Georgia, the Cherry Blossom capital of the world. 10 AM to 5 PM EST. For a Cherry Blossom certificate, send large SASE to Mona Wigherington, N4MPR, 2898 Williamson Drive, Lizella, GA 31052.

1986 marks the 50th anniversary of the Greater Cincinnati Amateur Radio Association. A number of special events are planned. Watch for announcements here.

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THE GUERRI REPORT by Ernie Guerri, W6MGI

millimeter waves part II: applications

Last month I discussed the domain of millimeter waves and some of the components used in this emerging part of the spectrum. This month we'll look at a variety of applications and see why there's so much interest, in professional circles, in moving into the mm wave region.

Military applications continue to lead the way in mm wave activity. Of special value to the military is the fact that mm waves are more difficult to intercept than microwaves because of the very narrow beams of the antennas used at mm-wave frequencies. This advantage, combined with considerable spectrum available for frequency-hopping, makes for LPI (Low Probability of Intercept). The fact that relatively low power is used for many current mm wave applications is actually a benefit because it makes the unwanted detection of signals even more difficult. It should be noted, however, that special tubes that can deliver 500kW of peak pulse power at 100GHz are available!

Military applications should be of particular interest over the next several years, as high-tech countries gain significant advantage in key electronic areas by virtue of their ability to implement mm wave technology.

commercial, scientific applications

The principal commercial applications for mm waves will be the same as for microwaves — telecommunications and data transmission. The mm region opens some significant opportunities in very wideband data/video transmission that were difficult to accomplish in the microwave region because of spectrum crowding and antenna beamwidth considerations. As computer graphics become an essential part of business communications, the relatively wide bandwidths needed for animated graphics can easily be accommodated in the region above 30 GHz. We can also expect that the fidelity with which complex waveforms can be transmitted in the mm region will give rise to new coding techniques that will permit more efficient spectrum utilization, lower error rates, and greater data security than has previously been possible. And — as if we weren't already up to our ears in data — there will be a vast expansion of telephone, FAX, and real-time video capacity with simpler, smaller terminals.

The management of Earth's resources and related scientific applications will also benefit from developments in the mm region. Experiments have already confirmed that mm-wave radar, combined with synthetic aperture techniques, can yeild details of the Earth's surface, and sub-surface, that weren't even detectable with microwave techniques.

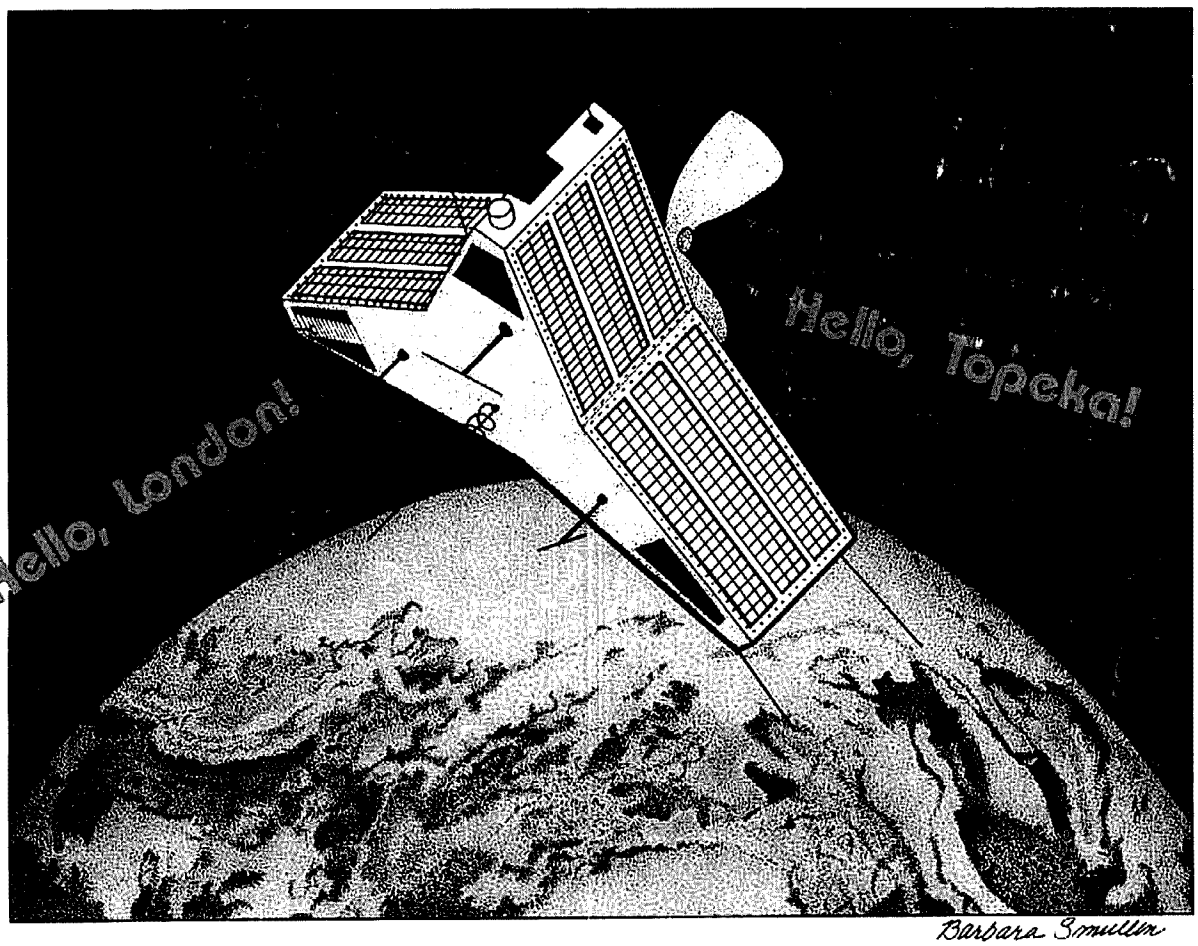
One scientific domain that will reap special benefit is deep space radio-astronomy. Because the Earth's atmosphere attenuates so much of the mm region, very little data on intergalactic radiation in this part of the spectrum exists. As more mm equipment becomes available for Space Shuttle missions or orbital laboratories, a whole new "picture" of the structure of space, and of the physical mechanisms at work, is likely to emerge.

Another especially promising area for practical applications of mm wave technology is air transport safety. Because of the excellent spatial resolution and high doppler rates possible at radar frequencies above 30 GHz, small, lightweight aircraft collision-avoidance systems (CAS) will become more practical and accurate. CAS depends on knowing the position of potentially dangerous encounters with extreme accuracy and on having dependable information regarding the direction and closing rate of oncoming aircraft. These technical requirements are made to order for the characteristics of mm waves.

Last, but not least, are possibilities just now being observed in medical and biological research. Some researchers have noted that the very shallow tissue penetration of low-power mm waves can give a "radar" image of thermal activity, blood flow, and other physical and neurological functions that is not possible with more conventional techniques. This is very new work, and reliable data is being developed by only a handful of investigators. However, a few pioneers are wasting no time in exploring the prospect that this new RF domain can bring improved levels of health and well-being to all of us.

ham radio

ham radio



Barbara Smuller

hr

satellite communications

focus on communications technology

ham radio magazine

APRIL 1986

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T. H. Tenney, Jr., W1NLB
publisher

Rich Rosen, K2RR
editor-in-chief
and associate publisher

Dorothy Rosa, KA1LBO
assistant editor

Joseph J. Schroeder, W9JUV
Alfred Wilson, W6NIF
associate editors

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editorial production

editorial review board

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publishing staff

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Rally Dennis, KA1JWF
director of advertising sales

Dorothy Sargent, KA1ZK
advertising production manager

Susan Shorrock
circulation manager

Therese Bourgault
circulation

cover art:
Barbara Smullen

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REFLECTIONS

dual roles

When is an "amateur" not an "amateur?" One possible definition is when he applies his knowledge and uses it in a technical trade such as Engineering or Science. At the recent RF Technology Expo 86 (January 30-February 1), thousands of similarly interested technical individuals got together for three days in Los Angeles to discuss the latest developments in the RF communications field. The technical exhibition, sponsored by *rf design*, featured technical forums in which 79 papers on HF through microwave subjects were presented and an exhibition by 133 manufacturers and their representatives who displayed their wares, from crystal oscillators through interactive computer-aided design applications and software.

Just a brief scan of the titles of some of the papers presented at the show reveals topics of considerable interest to Radio Amateurs. For example, Eyring Research Institute showed the proper way to evaluate HF antennas on a large scale. Believe me, their antenna "test bed," instrumentation, and procedures had many in the audience on the edge of their chairs, taking in every word for possible use — on a more limited basis, perhaps — back at their own QTH. Actually, many of the engineers who remained after the end of this session in order to dig for more facts turned out to be Radio Amateurs who happened to be engineers as well. It was difficult to tell which aspect of their experience elicited more questions — the "amateur" or "commercial."

Which brings me to my main point. The engineers and scientists at the show were, in many cases, Radio Amateurs who, over the years, had been able to combine their interest and avocation to the mutual benefit of both. That a connection exists was further indicated by several speakers who quite independently mentioned that they find both *rf design* and *ham radio* good sources of HF communications information.

"But wait a second," you say, "what do our interests have in common with the topics discussed at the show? To answer that question, take a look at this abbreviated list of some of the topics covered:

Choosing the right crystal and oscillator
High efficiency power amplifiers
IM, phase noise, and receiver dynamic range
High-pass filter design
Increasing the bandwidth of helical antennas
How to make simple test equipment
New low-power SSB circuits
Designing combline and interdigital bandpass filters
How to bias RF and microwave transistors
Wideband modules using FETs
Practical wideband RF power transformers, splitters, and combiners
RF power amplifier design
Understanding RF transistor data sheets
1-kW, solid-state L-band amplifier (What about you 1296 fans?)
Broadband HF antenna testing
ACSSB and SSB communication receiver testing
RF circuit design using interactive computer-aided graphics
Wideband high dynamic range front-ends
High-Q inductors using powdered iron cores

If you're interested in reading any of the 79 papers, the complete set* has been bound into a 2-inch thick compendium that weighs in at approximately 4 pounds. (I should know, I carried mine around with me for the entire three days.)

If you find some of these "engineering" topics of special interest, let me know . . . perhaps some of the authors would be interested in writing for *ham radio*.

I truly believe that as a result of intense interest and hard work on the part of many Radio Amateurs, great strides have been made in the most technically demanding fields of communications and will continue to be made by those individuals sharing this dual role.

Rich Rosen, K2RR
Editor-in-Chief

*For information, contact Cardiff Publishing, 6530 South Yosemite Street, Englewood, Colorado 80111 (303-694-1522).

presstop de W9JUV

SIGNIFICANT IMPACT ON THE "ELECTRONIC COMMUNICATIONS PRIVACY ACT" appeared likely as a result of the January 30 House subcommittee meetings in Washington. Strongly supporting the bill were two spokesmen for the telephone system and a Tandy representative -- though Tandy's speaker was teamed with ARRL and the Association of North American Radio Clubs (ANARC) in the apparent belief Tandy would oppose the bill as scanner supporters. However, Tandy came out in favor of it from their position as a cellular telephone supplier.

ARRL Shifted From Its Previous Position That Exempting Amateur Radio satisfied League concerns; Perry Williams, W1UED, while approving the exemption, pointed out that Amateurs -- and others, as well -- have many legitimate reasons to listen across the radio spectrum and the bill would make many such activities illegal. Probably the most telling testimony came from ANARC's Terry Colgan, WD5GWC, who not only pointed out various fallacies in the bill when applied to radio communications but demonstrated how effective and inexpensive available encryption devices are. (An article on the hearing will appear soon in Ham Radio.)

THE COMIC BOOK PROMOTION FOR AMATEUR RADIO being funded by the Amateur Radio industry is moving along well, ARRL's Dave Sumner, K1ZZ, reported at a February 7 meeting during the Miami Tropical Hamboree. The group decided to proceed with a story line based on the popular "Archie and His Friends" strip, with final approval in the near future.

Lack of Well-Qualified Amateur Radio Instructors is a major problem in effective training and growth, Gordon West, WB6AOA, told the group. He proposes a program, possibly through the ARRL, to promote instructor training. Dealer involvement in Amateur Radio promotion was also considered. The next industry group meeting is set for April 24 in Dayton.

MODULATED CW IS NOW PERMITTED ON 10 METERS, but only from 29.5-29.7 MHz. Acting on PR Docket 85-168 at its February 19 meeting, the FCC authorized F2A emission on the band's top portion in order to enable repeaters to identify using Morse code (effective date: April 23).

THE "PACIFIC AREA COORDINATION ASSOCIATION" is a newly formed regional VHF-UHF effort to promote wide-area coordination. Organizer WA6DPJ has sent invitations to coordinators west of the Continental Divide; send SASE to Box 23183, Pleasant Hill, CA 94523 for details.

Mississippi Will Retain 15 kHz Spacing On 2 Meters' top end. Southern California is now the only area of the country to use inverted 15 kHz splits on 2 meters, following a shift to "upright" splits by repeaters in western Colorado.

Northern California Could Shift To 20 kHz Spacing on 146-148 MHz and still accomodate all existing repeaters in its area, the Northern California Relay Council (NARC) reports. The plan will be considered at NARC's April 5 meeting in Sacramento.

NARC Has Also Proposed A 33-cm Band Plan that is essentially identical to the ARRL's interim plan developed by the VUAC. A push seems to be building within the VUAC to come up with a satisfactory final plan for the new 902-928 MHz band.

Repeater Coordination And Spectrum Management Will Be The Subjects of an on-going workshop during the Dayton Hamvention weekend. Location and other details of this crucial activity may not make the printed program, so check with Hamvention officials upon arrival.

REPLACEMENT OF THE GMRS WITH A NEW "CONSUMER RADIO SERVICE" was proposed by the FCC in a Notice of Inquiry adopted January 30. A total surprise to the GMRS's sophisticated and well organized users, the proposal apparently stemmed from a synthesis of the GMRS's own proposal to expand its utility through new technologies and from the Commission's desire to establish a new short-range quality radio service for the general public. After previous efforts to find spectrum for such a service in the 900-MHz band fell flat, the FCC apparently decided that GMRS's two 200-kHz slots in the 460 MHz band was an acceptable alternative.

"Personal Directed Communications" Is The Apparent Direction of the FCC's thinking, employing "user transparent" sophisticated portable equipment "designed to control users' actions automatically." They also ask whether "one-way" (paging) should be included, and for suggestions as to how much and what kind of automation should be included.

GMRS Licensees And Users Are Extremely Upset, and understandably so, with their well established system of repeaters, mobiles, and portables -- very active in personal, business and public service communications -- threatened with extinction.

Comments On PR Docket 86-38 Are Due At The FCC May 30, and the Reply Comments June 30.

User-Programmable Land Mobile Radios May Be Banned as a result of a Notice of Proposed Rule Making approved by the Commissioners at their January 30 meeting. In response to complaints of interference from radios reprogrammed to unauthorized frequencies, PR Docket 86-37 would prohibit the FCC from type accepting Part 90 radios operating above 25 MHz that have external frequency control. Though it's known at the FCC that many synthesized Amateur radios are reprogrammable to non-Amateur frequencies, Part 97 equipment wasn't included.

21 SCHOLARSHIPS FOR STUDENTS WITH GENERAL OR HIGHER LICENSES are available from the Foundation For Amateur Radio. Write FAR, 6903 Rhode Island Ave., College Park, MD 20740.



comments

spreadsheets for EME

Dear HR:

I enjoyed KE6ZE's informative article on EME ("EME-link Calculator Program," February, page 70). By a very odd coincidence I used the same equation from the *ARRL Handbook* to illustrate the convenience of spreadsheet programs for difficult calculations in an article in the same month's *QST*.

I hope that those readers who can compare David Engle's program written in BASIC with my speadsheet template will comment on the relative merits of the two approaches. It is, I think, important for hams to show leadership in matters of this sort which test techniques of immediate importance for technology.

My prejudice, of course, is that the spreadsheet is faster, allows for easier correction of errors, and has more versatility in printing results, than programs written in traditional programming languages.

I would be pleased to hear your readers' opinions.

Dick Ward, KC8OH
East Detroit, Michigan

understanding telephones

Dear HR:

Thank you for the many articles such as "Understanding Telephones" (by Julian Macassey, N6ARE), which appeared in the September issue.

Please do not assume we should know it. *Print it.*

Owen Zweiger, KD7WL
McMinnville, Oregon

propagation disks available

Dear HR:

Response to my VHF propagation articles (July, 1985 and January, 1986) has been surprising, with dozens of hams requesting my program on disk for the Commodore 64. Because of requests from owners of other computers, I now have a working version ready for the IBM-PC.

I will supply a disk with the complete program (VHF through L-Band) in an IBM-PC format for $8.00.

Lynn Gerig, WA9GFR
Route 1
Monroeville, Indiana 46773

cable comments

Dear HR:

Joe Reisert's column, "VHF/UHF World," in the October, 1985, issue of *ham radio*, is one of the best synopses on coaxial cable that I have read in Amateur publications, and, for that matter, in industry publications in some time. My sincerest compliments and congratulations. It was obvious that W1JR made some extra efforts in trying to document and assemble the information.

Two comments that I should like to make with respect to his article are not a reflection of my current job description and/or position. Rather, they are a reflection of my past experience as a product engineer with specific responsibilities of developing and obtaining MIL-SPEC qualification on coaxial cables.

On page 89, Joe indicates avoiding the contaminating types of jackets at all costs.

Yes, I agree with his basic position ("Don't be penny-wise and pound-foolish") but at the same time, his position does cost the average Amateur Radio Operator some unnecessary dollar expenditures. The difference between contaminating and non-contaminating jackets is the amount of plasticizer in the vinyl compound, which then provides flexibility at low temperature extremes. It has been my experience in laboratory environments with so-called contaminating jackets that circulating ovens of 120 degree C (which is more than adequate temperature to drive out plasticizers) after seven days shows only a minimal amount of plasticizer migration. Outside of the laboratory environment, I personally have not experienced plasticizer migration problems with coax, both buried and fully exposed to the elements. After eight to 10 years of service, no attenuation increases.

At the same time, I have examined many pieces of coaxial cable in which plasticizer migration and/or contamination has been submitted as the cause of high attenuation. To date I have not seen a piece of coaxial cable in which that has been the cause of high attenuation in real-life operation. It has always been water and moisture getting into the cable. Water and moisture entry most commonly come about from inadequate sealing at the connector ends of the cable and/or cuts or pin holes caused by abrasion to the jacket. Yes, I see W1JR's point — and I have heard the myth many, many times throughout my active days as an engineer and in my current position as well.

One other small picky comment is in the last part of his article on page 91. He indicates that CATV transmission line is typically specified up to 350 MHz. This typical specification has to do only with structural return loss, which is a test for periodicity, which causes frequency suck-out. His advice is quite proper, then, to test it at the frequency of interest before installing it, but specifications up to 350 MHz have nothing to do with power handling capabilities and/or attenuation characteristics.

Once again, as an active and involved Amateur Radio Operator for better than 25 years, a tip of my hat for a fine article.

Ronald L. Steir, W9ICZ
Marketing Director
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Richmond, Indiana 47375

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Do you want to access a new Amateur band that's always open when it's supposed to be? A band that doesn't fade away without warning, that makes DX contacts sound like locals, and has no skip zones?

Listening to, or working through Amateur communications spacecraft isn't difficult, but most newcomers simply don't know how to go about it properly. Not sure of what they're doing, they usually achieve disappointing results; deciding that the amount of effort invested must be so much more than the results achieved, they give up and go back to their regular haunts, where they can usually at least find someone to talk to. This is a shame, because satellites have come of age and commercial equipment is as readily available for the satellite bands as for the regular HF or VHF bands. You can buy or roll your own, but in either case — just like on 20 meters or the other HF bands — you have to have some knowledge of what's going on if you're going to get the maximum enjoyment out of the equipment.

terminology

A communications satellite is basically a repeater in the sky. It receives signals transmitted up from the ground on one Amateur band and retransmits the same signals down to the earth on a second Amateur band. It's part of a communications link between two Amateur stations on the ground as shown in **fig. 1**; signals on their way up to the satellite are said to be *uplinked* by stations on the ground while the corresponding signals coming down from the satellite are being *downlinked*. As the satellite orbits the earth it passes over different locations; the point immediately beneath the satellite at any time is called the *subsatellite point*.

The area of the earth's surface that the satellite can "see" depends on its altitude; the higher it is, the more it can see. A commercial communications satellite in a high altitude over the equator can see about one third of the earth's surface. A satellite at a low altitude sees much less.

Any station that the spacecraft can see, can see the spacecraft. When a station can see the spacecraft, it is said to be *in range*. Thus any two stations in range of the satellite at the same time are said to have a *window* into the satellite and can communicate through it.

Most orbits are elliptical rather than circular. The highest point above the surface of the earth in the orbit is called the *apogee*; the lowest point of that same orbit is the *perigee*.

Even though the orbit of the satellite is fixed, the earth rotates beneath it. The time it takes for the satellite to travel once around its orbit from the place where the sub-satellite point crosses the equator to the next time the sub-satellite point crosses the equator going in the same direction is called the *period* of the orbit. When the sub-satellite point has returned to the equator, the point on earth that was previously below it will have moved away because of the rotation of the earth; consequently, a new location will be beneath it. The number of degrees of longitude that have passed by during this time is known as the *orbital increment* (see **fig. 2**). The first orbit of the day is known as the *reference orbit*.

Earth stations will see different parts of different orbits as shown in **fig. 3**. The azimuth, or horizontal bearing and elevation angle to the spacecraft, will change with the orbit. The spacecraft will appear to rise above the horizon when it enters the range of the ground station. The time at which the spacecraft rises above the horizon is called *Acquisition Of Signals*, or

By Joe Kasser, G3ZCZ, P.O. Box 3419, Silver Spring, Maryland 20901

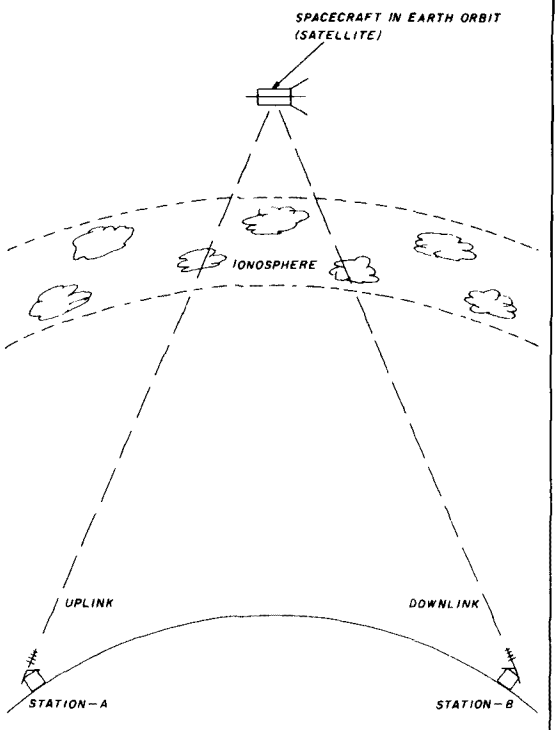


fig. 1. Satellite communications path. Station A is transmitting (uplink). Signals are received by the satellite and re-transmitted down to station B (downlink).

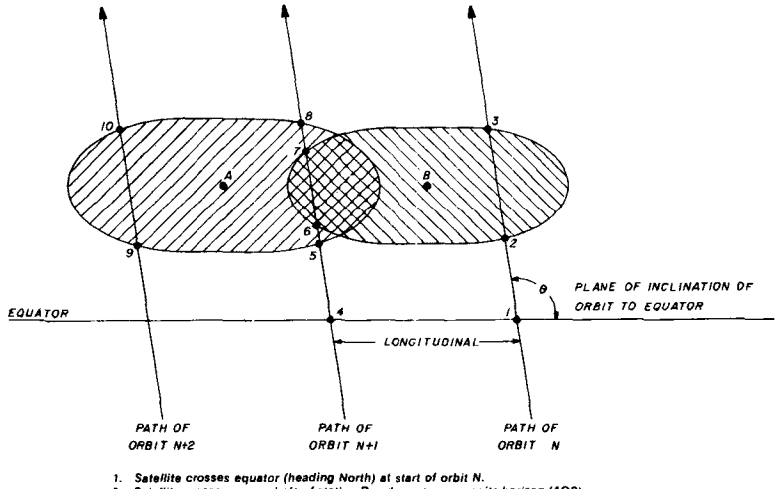
AOS. The position of the satellite in the sky as seen by the ground observer will change as it passes along its orbit, rising higher and passing across the sky, getting lower, and then finally setting on the horizon. The time at which it sets beneath the horizon of the ground station is known as *Loss of Signals*, or *LOS*.

The path traced by a satellite in the sky as seen by a particular ground station will vary according to the type of orbit. The path traced by a satellite in a circular orbit will usually approximate a section, or chord, of a circle. The path traced by a satellite in an elliptical orbit will depend on the apogee and perigee of the orbit and how close the observer is to the subsatellite point.

characteristics of satellite signals

In order to copy signals from satellites, we first need to know a little about the types of signals we're trying to receive. At any particular time, an observer on the ground may see the satellite in any direction with respect to the horizon (*azimuth*) and at any altitude between the horizon and a point directly overhead (*elevation*). This means that signals from various satellites arrive at a receiving station from any angle in any direction.

Radio waves are generated in a polarized manner. Conventional Amateur station antennas may generate vertically or horizontally polarized signals, depending on the position of the antenna with respect to the



1. Satellite crosses equator (heading North) at start of orbit N.
2. Satellite crosses range circle of station B and comes up over its horizon (AOS).
3. Satellite crosses range circle of station B and sets on its horizon (LOS).
4. Satellite crosses equator at start of orbit N + 1.
5. Satellite crosses range circle of station A (AOS).
6. Satellite crosses range circle of station B (mutual access window opens).
7. Satellite crosses range circle of station B (mutual access window closes).
8. Satellite crosses range circle of station A (LOS).
9. Satellite crosses range circle of station A on orbit N + 2 (AOS).
10. Satellite crosses range circle of station A on orbit N + 2 (LOS).

fig. 2. Looking down on the earth from the satellite showing track of spacecraft on surface (subsatellite point). Range circles are drawn as ellipses on rectangular (map) projection. Circles overlap to give mutual access window on orbit N + 1 between points 6 and 7.

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TYPICAL SATELLITE ANGLES FOR ONE GROUND LOCATION ON ONE DAY

OSCAR-9

U.T.C. HHMM:SS	AZIMUTH DEGREES	ELEVATION DEGREES
1330:00	102	10
1332:00	68	14
1334:00	35	9
1336:00	16	2
1500:00	194	1
1502:00	207	10

OSCAR-10

U.T.C. HHMM:SS	AZIMUTH DEGREES	ELEVATION DEGREES
1600:00	256	7
1700:00	240	12
1800:00	236	10
1900:00	235	6
2000:00	236	1

RS-5

U.T.C. HHMM:SS	AZIMUTH DEGREES	ELEVATION DEGREES
1748:00	4	3
1750:00	8	9
1752:00	15	17
1754:00	24	25
1756:00	39	34
1758:00	62	41
1800:00	91	42
1802:00	115	35
1804:00	131	26
1806:00	140	17
1808:00	147	10
1810:00	151	3

OSCAR-11

U.T.C. HHMM:SS	AZIMUTH DEGREES	ELEVATION DEGREES
0824:00	359	5
0826:00	350	15
0828:00	327	31
0830:00	269	38
0832:00	231	21
0834:00	218	9

RS-7

U.T.C. HHMM:SS	AZIMUTH DEGREES	ELEVATION DEGREES
1856:00	347	5
1858:00	344	12
1900:00	340	21
1902:00	333	32
1904:00	318	45
1906:00	286	56
1908:00	242	54
1910:00	217	41
1912:00	205	28
1914:00	199	18
1916:00	195	10
1918:00	192	3

Space shuttle Challenger

U.T.C. HHMM:SS	AZIMUTH DEGREES	ELEVATION DEGREES
0625:00	328	0
0626:00	330	5
0627:00	334	10
0628:00	343	20
0629:00	7	37
0630:00	77	43
0631:00	113	24
0632:00	124	13

Shown are azimuth and elevation angles from the ground station to the different spacecraft at different times of the day. It can be seen that in order to adequately copy signals from the spacecraft, the ground station must be able to receive signals coming from any azimuth or elevation. This figure only lists data for one of the daily passes for each satellite. As a rule they will be audible at other times of the day with signals coming from other directions.

fig. 3. Typical satellite azimuth and elevation angles for various passes as seen by a representative ground station.

ground. If the radiating elements are horizontal, the antenna is said to be generating horizontally polarized signals; conversely, if the elements are vertical, the antenna is vertically polarized. The same polarization also holds for reception. Thus, vertical antennas receive vertically polarized signals best and horizontal antennas receive horizontally polarized signals best. True vertically polarized antennas will copy little or no horizontally polarized signals. Two-meter and other VHF/UHF FM antennas are vertically polarized, while base stations working SSB/CW use horizontal antennas. This is because automobile antennas are vertically polarized, and the mobile stations put weak signals into horizontal antennas. (In the early days of mobile radio communications, Amateurs fitted "halo" antennas on their cars to send and receive horizontally polarized signals in order to be compatible with the base stations. When the mobiles using FM began to outnumber

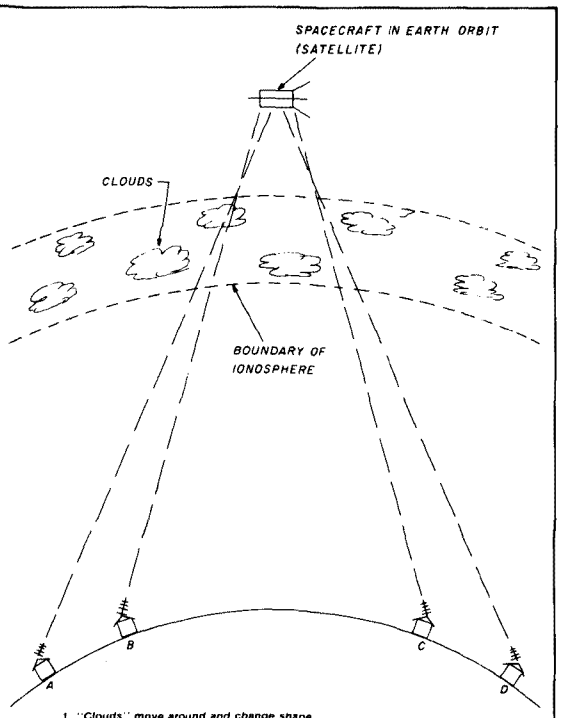


fig. 4. Some factors affecting satellite communications.

the fixed stations, there was no further need to use horizontal polarization and verticals became the rule. Nowadays, any base station that wants to use FM has to use vertical polarization.

On the HF bands both types of antennas are used interchangeably and everyone manages to work everybody. This is because the polarization of the radio waves changes as the signals pass through the ionosphere. A process known as *Faraday rotation* rotates the polarization of the signals. The signal as received on the ground is not entirely vertically or horizontally polarized and as such may be copied at somewhat lower signal strength on any antenna. Perhaps the good performance of quad antennas is due to their having both vertical and horizontal elements. When conditions in the ionosphere are changing, the received signals may appear to fade — i.e., get weaker and stronger as the plane of polarization is rotated by the ionosphere.

Satellite orbits are outside the ionosphere, which means that signals from the spacecraft are affected by the ionosphere in a manner similar to that which affects conventional terrestrial signals: the polarization of their signals changes. Conventional contacts tend to use the same part of the ionosphere. The ionosphere is not a constant layer above the earth, of course, but is instead made up of patches, or clouds. Since the satellite is moving, its uplink and downlink signals will pass through different parts of the ionosphere at different times, and the effects of the ionosphere on the signals will differ as time passes, as shown in **fig. 4**.

Not only does the ionosphere refract radio waves and change their polarization, it may also attenuate signals or even absorb them. As the spacecraft travels along its orbit, it may be spinning or tumbling, or the satellite itself may shield the on-board antenna from the receiving station. Because of the limitations of its equipment, the transmitter on the space vehicle is transmitting at a relatively low power — less than 10 watts output. Consequently, signals from satellites may arrive at the ground from any direction in azimuth or elevation, with any polarization, and at any signal strength (usually very weak). All these may, and usually do, vary as a function of time.

an ideal satellite receiving antenna

The ideal antenna for copying satellite signals should be rotatable in azimuth and elevation in order to cope with all the possible directions from which signals may arrive. It must be immune to changes in polarization if it is to cope with horizontal, vertical, and in-between polarization caused by Faraday rotation in the ionosphere. It must also have a reasonable amount of gain in order to cope with the fading in the already weak signals generated at the satellite.

Vertical and horizontal polarization are two kinds of linear polarization. Radio signals can also be circularly polarized. A circularly polarized antenna will respond equally to horizontally or vertically polarized signals — that is, changes in the plane of polarization will not be detected. Circular polarization also comes in two kinds, left-hand and right-hand (clockwise and counter-clockwise). To compound the problem, lefthand circularly polarized signals are not well received on righthand circularly polarized antennas and vice-versa.

antennas in common use on 10 meters

Figure 5 lists the commonly used bands in the Amateur Satellite Service. The most commonly used downlink bands are 10 meters, 2 meters and 70 cm. The first band combination that most people try is the 10-meter downlink and the 2-meter uplink commonly known as *Mode A*. This is because they usually have 10-meter capability in their stations and can thus attempt to copy the satellite without adding too much equipment.



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 2 METERS       145.8   -  146.00 MHz
70 CENTIMETES   435.0   -  438.00 MHz

2. SATELLITE TRANSPONDERS

SPACECRAFT      MODE      UPLINK                      DOWNLINK
-------------------------------------------------------------------------
AMSAT - OSCAR 6   A    145.85 - 145.95MHz          29.45 - 29.55MHz
AMSAT - OSCAR 7   A    145.85 - 145.95MHz          29.40 - 29.50MHz
AMSAT - OSCAR 7   B    432.125 - 432.175MHz        145.975 - 145.925MHz
AMSAT - OSCAR 8   A    145.85 - 145.95MHz          29.40 - 29.50MHz
AMSAT - OSCAR 8   J    145.90 - 146.00MHz          435.10 - 435.20MHz
 -------------------------------------------------------------------------
The data supplied on AMSAT - OSCARs 6 - 8 is for historic purposes as
the spacecraft are no longer operational.

CURRENTLY ACTIVE

AMSAT - OSCAR 10  B    435.05 - 435.15MHz          145.95 - 145.85MHz
AMSAT - OSCAR 10  L    1269.05 - 1269.85MHz        436.95 - 436.15MHz
RS - 5            A    145.91 - 145.95MHz          29.41 - 29.45MHz
RS - 7            A    145.96 - 146.00MHz          29.46 - 29.50MHz
-------------------------------------------------------------------------

FUTURE (PROPOSED) SPACECRAFT

For launch in early 1986
RS - 9/10         A    145.96 - 146.00MHz          29.46 - 29.50MHz
RS - 9/10         K    21.26 - 21.30MHz            29.46 - 29.50MHz
RS - 9/10         ?    21.26 - 21.30MHz            145.96 - 146.00MHz

The RS spacecraft have been ground tested and are due for  launch
in 1986.

FUJI - 1          A    145.85 - 145.95MHz          29.40 - 29.50MHz
FUJI - 1          M    1267.55 - 1267.75MHz        436.00 - 435.80MHz

The FUJI spacecraft is being built in Japan under the control  of
JAMSAT, a group of Japanese Radio Amateurs.

ARSENE

The  ARSENE spacecraft built by a group of French Radio  Amateurs
is  supposed  to be launched in the demonstration flight  of  the
ARIANE 4 rocket in 1986.  It will contain a Mode B transponder.

AMSAT -PHASE 3C

The AMSAT Phase 3C spacecraft is also scheduled for launch in mid
1986.   It  will  contain a Mode B transponder as well  as  other
transponders  having either uplink or downlink capability on  the
higher  frequency bands.

AMSAT  has a policy of not obsoleting user equipment,  so mode  B
will  be  around  for a long time.  As mode A  is  an  excellent
introductory  mode,  it  can be expected on any  further  general
purpose  Phase  2  type spacecraft.  The Russians also  tend  to
favour hf so mode A and possibly mode K will also be around for a
while.
```

fig. 5. Commonly used satellite communications bands.

Once you're hooked on receiving, the price of a transmitter usually becomes a justifiable expense. Although putting together a minimal receiving and transmitting station isn't difficult, steerable antennas for the 10-meter band are relatively large. Therefore relatively few Amateurs can steer their 10-meter antennas in both azimuth and elevation. Steerable antennas for 2-meters and 70 cm are much smaller and as a result, more manageable.

Antennas in common use on the 10-meter band include verticals and multi-element beams optimized for working DX. As such, they have very good responses to signals arriving from low angles but are not at all suited for signals arriving at high angles. Vertical antennas respond to low-angle radiation from all directions, while beams respond to low-angle signals from the direction in

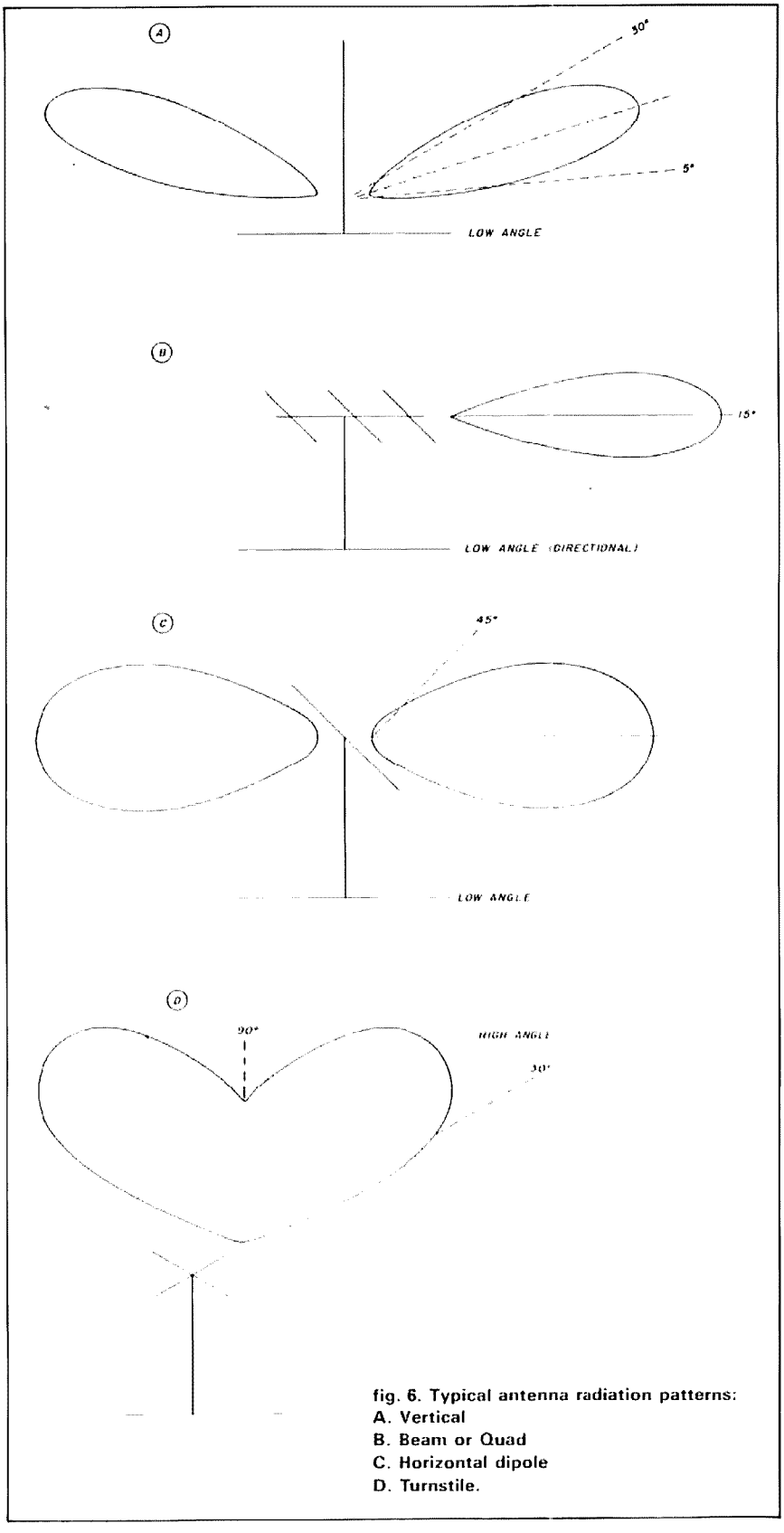


fig. 6. Typical antenna radiation patterns:
A. Vertical
B. Beam or Quad
C. Horizontal dipole
D. Turnstile.

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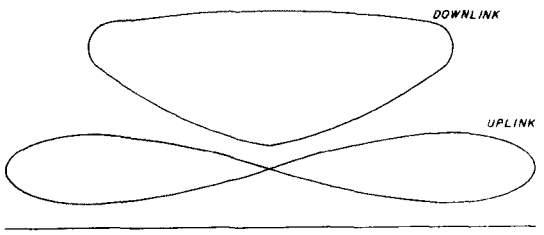


fig. 7. Typical VHF/UHF operator antenna characteristics. Hears well when satellite is high in sky. Talks well when satellite is low in sky.

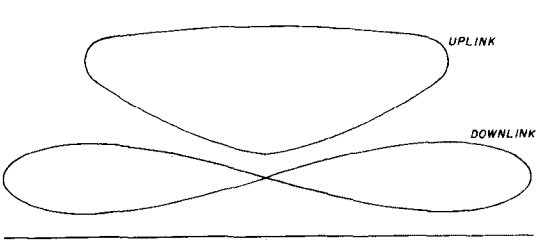


fig. 8. Typical HF operator antenna characteristics. Hears well when satellite is low in sky. Talks well when it is high in sky.

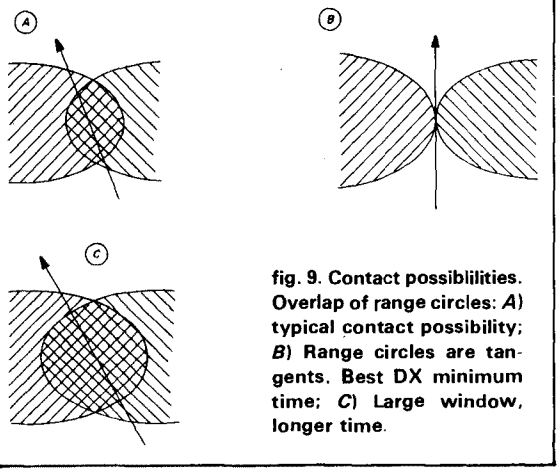


fig. 9. Contact possiblilities. Overlap of range circles: *A*) typical contact possibility; *B*) Range circles are tangents. Best DX minimum time; *C*) Large window, longer time.

which they happen to be pointed. Stations using these antennas have trouble hearing signals arriving from higher angles.

Conventional literature has touted the turnstile, or crossed dipole antenna, as the answer to the problems of satellite reception at 10 meters. It has circular polarization and a high-angle response pattern. It does very well when the satellite is located at elevations greater than about 30 degrees as seen by the observer, but has a poor response to signals arriving at low angles (close to the horizon). Typical radiation patterns for these antennas are shown in **fig. 6**.

Most Amateurs who have problems working Mode A fall into one of two categories. The first category includes the VHF/UHF operator who decides that satellites offer both a technical challenge and increased opportunity for some exciting DX. This operator usually has excellent linear (horizontal or vertical) polarized antennas for the 2-meter uplink bands but has nothing for 10 meters. Reading that a turnstile can be a simple, effective device for reception, he builds one and finds that, sure enough, he can hear something. It may be weak, but, by golly, those signals are coming from outer space!

Step back for a minute and analyze this situation as sketched in **fig. 7**. The uplink antennas on 2 meters can put a powerful signal into the satellite when it's at low angles of elevation as seen by this operator. His downlink antenna, however, receives best when the satellite is at a high angle. In other words, when he can hear it, he can't access it . . . and when he can access it, he can't hear it — meaning, he cannot hear himself.

The second type of Amateur who decides to have a go at satellite operation is the HF operator, who usually has a good beam antenna for 10 meters. Reading that a turnstile antenna is a good choice for satellite operation, he builds one and uses it. Now analyze this situation as sketched in **fig. 8**. The uplink antenna on 2 meters puts a weak signal into the satellite when it is at low angles of elevation as seen by this operator. His downlink antenna, however, receives best when the satellite is at low angles. In other words, when he can hear it, he cannot access it . . . and when he can access it, he cannot hear it — meaning that he cannot hear himself. Although this is the inverse situation to that of the VHF/UHF operator it has the same characteristics: both are "alligator operators" — all mouth and no ears.

There is a third category: the apartment dweller who cannot put up HF antennas at all. This type of operator can usually install some kind of VHF/UHF array on a balcony and work Mode B quite well. But when he tries Mode A, he has problems because of the size of his 10-meter receiving antenna.

It's no surprise, then, that the vast majority of Radio Amateurs who *decide to* become *active in satellites* have trouble working them at first.

matching uplink and downlink antennas

In order to get the most enjoyment out of satellite operation, it's necessary to match the uplink and downlink antennas. Before doing this, however, it's

necessary to consider other aspects of the satellite communications path.

The Earth-Satellite-Earth ccommunications link is a line-of-sight path. Each ground station has a range circle for which a window allows communications into the satellite. In order to work any other station, the range circles of the two stations must overlap as shown in **fig. 9**. The duration of any contact is governed by the time that the spacecraft spends in that window. Thus, the higher the elevation of the satellite as seen by the ground station, the shorter its communications range along the surface of the Earth. The best DX contact between any two stations occurs when the sub-satellite point of the orbit of the spacecraft passes over the ground where their range circles just touch — that is, at a tangent to both range circles. They will, however, also have very little time to make that contact.

antenna characteristics

The usual three-element Quad or Yagi-type antenna puts out a good directional low-angle signal. The turnstile antenna puts out a good omnidirectional high-angle signal. Vertical antennas put out good omnidirectinal low-angle signals. The 3/8 and 5/8 wave antennas used on 2-meters have good omnidirectional low-angle radiation characteristics. Somewhat directional high-angle radiation may be obtained from sloping dipoles attached between the top of a mast and the ground in the manner of guy wires (but don't ever use them as such), as shown in **fig. 10**. If you want to work the satellites successfully, you must match the characteristics of your uplink (transmitting) and downlink (receiving) antennas so that they have similiar radiation patterns.

receiving signals

The satellite downlink is usually marginal because the spacecraft is using low power and is far away. Every ESE (earth-satellite-earth) contact practically qualifies the spacecraft for yet another 1000-mile-per-watt award for QRP communications.

Most modern receivers (and others not so modern) suffer from a loss of sensitivity at the top end of the 10-meter band so that using a preamplifier to increase the strength of the received signals is a good idea. Most Amateurs feel that to communicate with DX stations they need the biggest antenna they can put up and the maximum power they can put out. But there's a fallacy at work in this kind of thinking; if the minimum amount of transmitted power to put an S-9 signal into a DX location is, for example, 100 watts, then for that transmitter to use 1000 watts would be a waste of power . . . or would it? For the moment, ignore the QRM factor in which the more power you use, the louder you are and the more likely you are to be heard

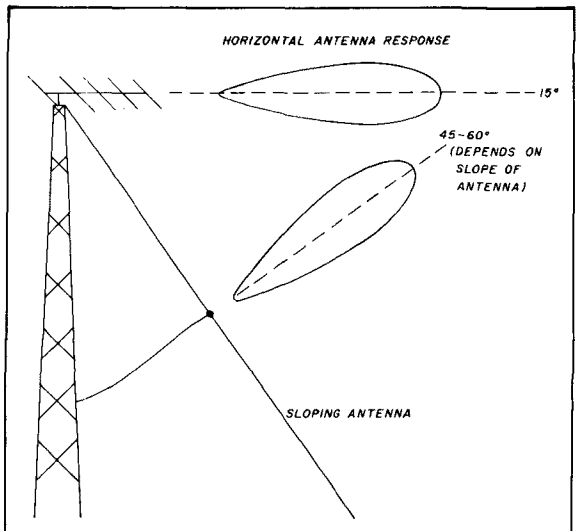


fig. 10. Sloping antennas for 10 meters. Optimally, put slopers on four sides of tower. These will work well for regular 10 meter contacts. Put dipole or beam on tower.

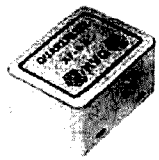
over the rest of the pack. If the signal is made weaker or attenuated by the ionosphere for one reason or another, what happens? In our example, we are receiving signals from a transmitter having the calculated 100 watts. If a fade equal to 5 S-units takes place, the received signal will drop down to S-4. This isn't too serious; S-4 signals can be copied, but what happens if the station is using the QRP and was S-4 to begin with? The same fade would take it down to S minus 1 or below the noise level, and no signals could be copied. The communications link should contain enough gain to minimize or avoid loss of reception due to extreme fading. In other words, some kind of margin should be built into the link.

the communications link

The communications link in a satellite contact can readily be split into two parts, the uplink and the downlink. Consider each of these in turn.

In the downlink, the transmitter output power is not under the control of the Radio Amateur, but is instead fixed by the satellite. The attenuation of the signals radiated by the satellite is a function of the distance between the spacecraft and the receiving station. The actual strength of the received signal at the ground station antenna will vary because of the attenuation due to fading and polarization changes in the ionosphere. Thus all the ground station operator can do is make sure that he has the best and most sensitive receiving capability that he can have. Ideally, the receiver should be such that the beacons on the downlink are receivable at good signal strength. In most

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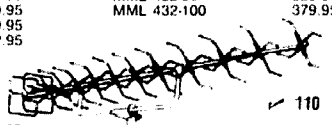
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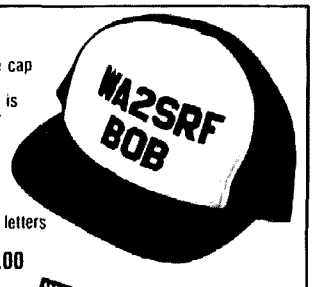
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cases, this means that a receiving preamplifier should be used ahead of the receiver.

In the uplink, the receiving antenna and on-board receiver sensitivity are governed by the design of the satellite. The attenuation of the signals from the ground as received by the satellite is a function of the distance from the spacecraft to the transmitting station. The actual strength of the received signal from the ground station antenna will vary because of the attenuation due to fading and polarization changes in the ionosphere. The effects of the ionosphere on the uplink may differ from those on the downlink. In the past, AMSAT has performed the link calculations before the launch of the spacecraft and released a recommended value in radiated uplink power (EIRP) for Amateurs to use with the satellite. This number has usually been conservative, and most satellite users have no trouble working through the transponder with much less power. The common solution to this problem is to boost the transmitter power until a good return signal is heard. This is not the optimal solution, because stations that have problems hearing themselves will tend to use too much power, not because they can't get into the satellite, but because they cannot hear themselves getting into it. The ionosphere may also behave differently in different places at any time, so that although the sending station may be having trouble hearing his own downlink, other stations further away may be copying him with ease. There's no easy solution to this situation. The compromise is to attempt to make your own signal as received on the downlink equal in strength to that of the transponder beacon. This means that you adjust your transmitter power to keep your own signal as strong as the beacon on your receiver. You can do this either by reducing the transmitter power gain or by aiming the antenna away from the spacecraft.

Gain in the communications link can be obtained by using amplifiers or directional antennas. Directional antennas are at a disadvantage in that they must be moved to track the satellite during the pass, while omnidirectional ones do not. On the other hand, they're usually cheaper than amplifiers, particularly high power UHF transmitting types. Thus, to obtain a certain power output level on the uplink, the Amateur has the choice of a directional antenna and low power or an omnidirectional antenna and high power, or something in between. Similarly, on the downlink, if the directional antennas are used, a receiving preamplifier may not be an absolute necessity. In any event, for reasonable results, make sure that the characteristics of your uplink and downlink antennas are matched.

locating the satellite

The common adage, "If you can't hear them, you

can't work them" must be modified for satellite users to read, "If you can't locate them, you won't hear them . . . and if you can't hear them, you can't work them."

In order to work satellites, an Amateur has to know not only when a satellite is in range, but also where to aim his antenna in order to put a signal into it. A number of different techniques have been developed over the past few years: graphical "circular slide rules" were first used very successfully for Phase 2 low-orbit satellites. As the personal computer found its way into ham stations, computer programs were developed to locate the satellites and the graphical plotters could be used to augment computer-generated data.

Fortunately, the first OSCAR satellites used by large numbers of Amateurs (AMSAT's OSCAR 6, 7, and 8 and the early RADIO spacecraft) were in circular orbits, which made locating them easy. All you had to do was pick a "reference orbit" as published in the Amateur Radio press and add the orbital increment to determine the position of the next equator crossing (start of the next orbit) and then add the period of the orbit to find the time of the following orbit.

When the first Phase 3 satellite (AMSAT's OSCAR 10) was put into service, it was placed into an elliptical orbit with a high apogee and a low perigee — definitely a non-circular orbit. AMSAT's Tom Clark, W3IWI, an astronomer by profession, wrote a program that utilized Keplerian elements for keeping track of the position of any satellite in the Amateur Satellite Service. This and many other programs have been widely disseminated and there should be at least one member of each radio club who knows how to get hold of them. (AMSAT can supply copies of such software through its Software Exchange.) Locating the satellite, therefore, should not be a problem.

reference

1. *Tom Clark*, W3IWI, "Basic Orbits," *Orbit*, March/April, 1981.

ham radio

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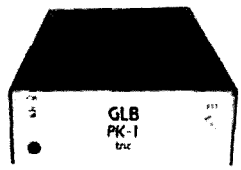
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Measure over
a 30 dB range
with confidence

wide-range RF power meter

Some time ago I decided to build a small antenna range. One of the key items I knew I'd need was an RF power meter with good stability and wide range. Most commercial units I found were beyond repair or the limits of my budget, and the homebrewed units were either limited in range or used modulation to avoid a drift problem.

I had used an LM11 operational amplifier in designing an earlier project and a friend later introduced me to an even better one. Some of the new chips coming on the market offer unbelievable performance and are slowly making system designers out of us circuit designers. A chip here, a chip there, follow the spec sheet as to optimum feeding — and we have a piece of test equipment that rivals commercial units.

I combined some of these into an RF power meter that features a 30 dB (useful to 35 dB) range from −15 dBm to −45 dBm, remote control, and good temperature stability. Although the antenna range is still in the future, the power meter has been used on the bench for evaluating hybrid couplers, helical filters, cavity filters, IF amplifiers, and such. I plan to use the power meter on the 70 cm band. But it can also be used from the HF band up into the GHz range.

theory of operation

The heart of the unit is the Hewlett-Packard HSCH-3486 zero-bias Schottky diode used as the detector. This device offers high voltage sensitivity and doesn't need the biasing featured in other detection schemes. The response curve is logarithmic from −50 dBm to −20 dBm; above −20 dBm the diode becomes increasingly nonlinear in detection response. The lower end is limited by the amplifier used.

To avoid using a modulation method of detection, a chopper stabilized operational amplifier was used. (The schematic is shown in **fig. 1**). The Intersil ICL7650 features an extremely low input offset voltage of 1 μvolt over a wide temperature range. The chopper op-amp basically converts the input DC voltage to AC, amplifies it, and converts it back to DC. Amplifying the DC output from the detector 150 times with a chopper op-amp puts the signal at a level that simpler op-amps such as the LM11 can handle. The National Semiconductor LM11 is a precision DC amplifier that combines the best features of existing bipolar and FET op-amps. Offset voltage is 100 μvolts and drift is 1 μv/°C. Six ranges in 5 dB steps are accomplished by this circuit by changing the gain of the amplifier. Each range is controlled remotely by reed relays. Offset voltages in the amplifier are nulled with two pots, one for the high range and one for the lower three ranges. These three devices — a diode which converts RF power into a logarithmic output equal to a dB scale and a pair of operational amplifiers — amplify AC microvolt level signals to volt levels, while introducing little drift.

construction

Originally the unit was to be mounted directly at the antenna and was therefore constructed in a diecast box for good shielding. Power is supplied remotely from a separate box, which also contains the meter and scale change (**fig. 2**). A schematic of the power supply is included in (**fig. 3**). When purchasing a dB scale meter make sure that the −3dB point falls exactly at half scale. Some meters have been "fudged" to accomodate circuit nonlinearities.

The inside of the box is shown in **fig. 4**. Its detection circuitry, visible on the left side, is shown in an enlarged view in **fig. 5**. The parts are mounted on a small piece of 0.015 inch brass shim stock and held in place by the TNC connector. Note the chip capacitor on the left, supporting the 50-ohm resistor. A value of 100 pF is adequate down to 10 MHz; below 10 MHz this value should be increased. For work above 70 cm up to 4.2 GHz, a coaxial-mounted detector is recom-

By Rudolf E. Six, KA8OBL, 30725 Tennessee, Roseville, Michigan 48066

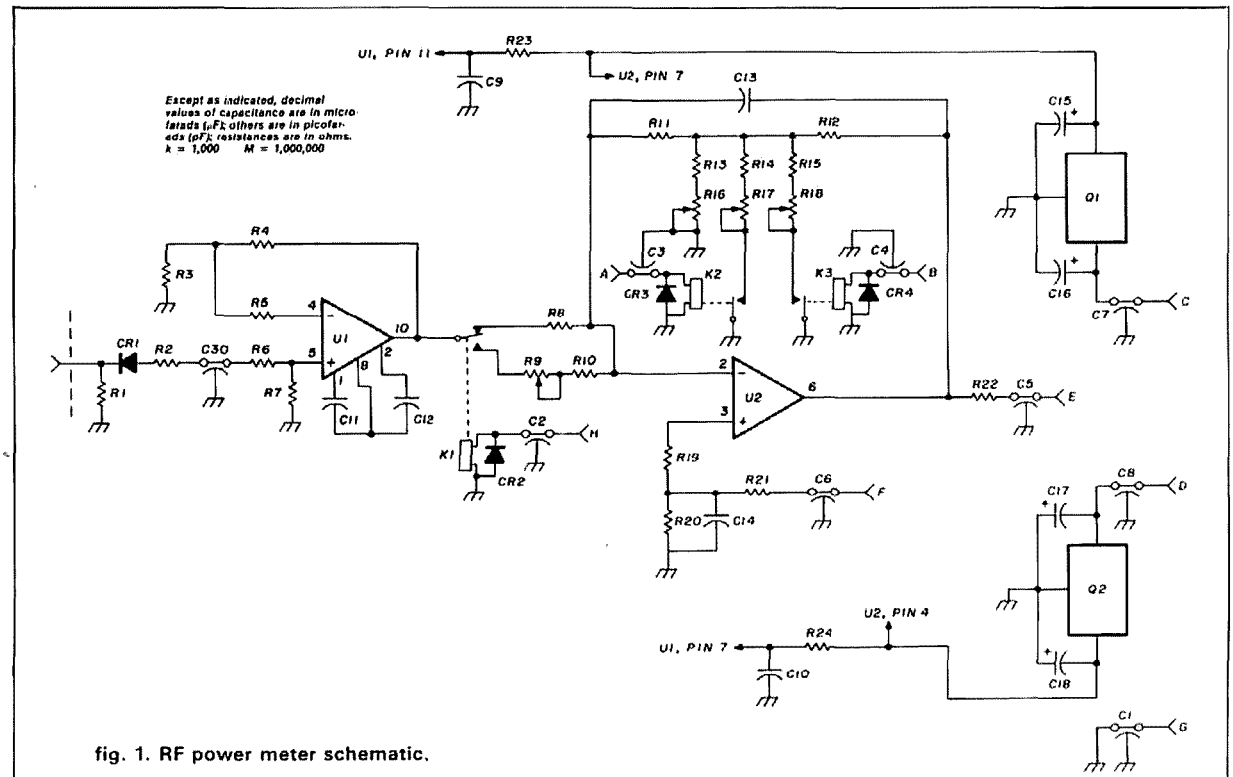


fig. 1. RF power meter schematic.

Table 1. RF power meter and power supply parts list

C1 thru C8	1000 pF feedthru (Erie, Cambion)
C9,10,15,18	1 μF 10wvdc tantalum
C11-12	0.1 μF metalized film
C13	500 pF disc
C14	0.01 μF disc ceramic
C16,17	2.2 μF 25 wvdc tantalum
C19,21	100 μF 15 wvdc electrolytic
C20	500 μF 15 wvdc electrolytic
C22,23	0.01 μF disc
C30	100pF chip capacitor
CR1	HSCH-3486 Hewlett-Packard
CR2,3,4,9,10	1N914 or equivalent
CR5,6,7,8	1N4003 or equivalent
K1	SPDT reed Magnecraft W172-DIP5 (internal diode — CR2 not used)
K2,3,4,5	SPST reed EAC EAC Z610-ND
M1	1 mA DC meter with dB scale
Q1,4	78L05 regulator
Q2	79L05 regulator
Q3	78L12 regulator
R1,2	50 ohm 1/8 watt carbon film
All resistors 1% metal film 1/4 watt	
R3,6,14,22	1k
R5,7	100k
R10	120k
R4	150k
R8,19	4.99k
R11,12	20k
R13	2.74k
R15	165 ohm
All resistors 5% carbon film 1/4 watt	
R20	100 ohm
R21	1 megohm
R23, 24	10 ohm
R27	1.5k
R9	50k Panasonic CEG54 trimpot
R16	500 ohm Panasonic CEG52 trimpot
R17	200 ohm Panasonic CEG22 trimpot
R18	100 ohm Panasonic CEG12 trimpot
R25,26	10k potentiometer
S1	DP6T rotary switch
T1,T2	6.3 VAC transformers
U1	ICL7650SCPD Intersil
U2	LM11CLH National
Box	CU-124 BUD
Chassis	9 1/2 x 5 x 2 chassis BUD Ac-403

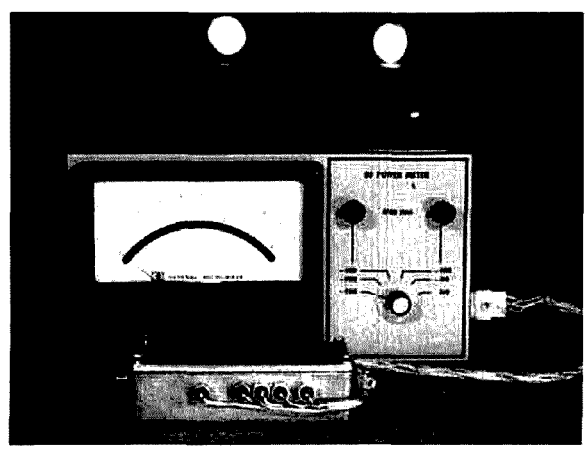


fig. 2. RF detector and amplifier mounted in a shielded enclosure. Range selection, meter and power supply are in a separate unit.

mended. A suitable unit, Model CD-51, is available from Elcom Systems Inc., 4032 Clint Moore Road, Boca Raton, Florida 33431-2895. The printed circuit board is suspended in the box (**fig. 6**). Two hangers made from 0.015-inch (0.038 cm) brass shim stock are soldered to the ground foil of the printed circuit board and are held by the feedthrough capacitors. Metal and

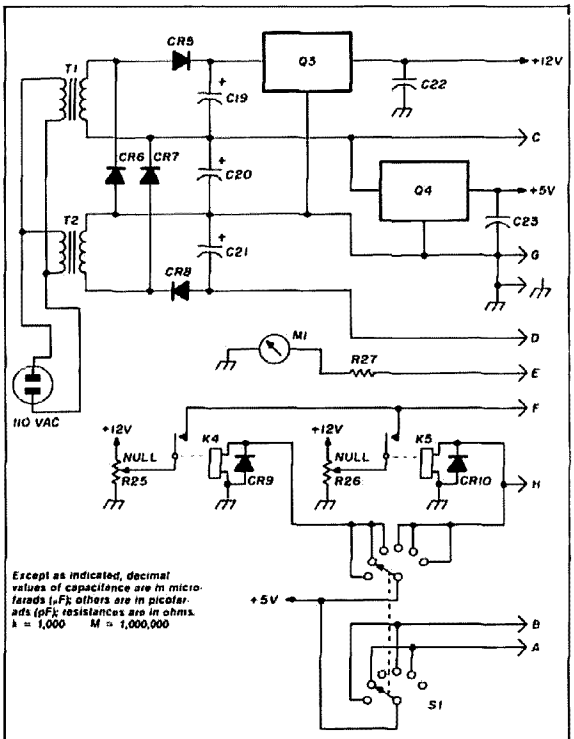


fig. 3. RF power meter power supply schematic.

carbon film resistors are used for accuracy and low noise. The PC board artwork and components layout are shown in **figs. 7** and **8**, respectively.

calibration

Calibration depends on the accuracy of the standard used. If you have no fixed attenuator, purchase the Model AT-51 5 dB TNC from Elcom ($14). Set the meter to the −15 dB range, check and adjust for zero with no signal applied. The meter zero pot has little control on this scale and if the meter doesn't read zero, there's something wrong with the circuit. Adjust a signal generator for a +30 dBm output level and turn R16 for full scale or 0 dB on the meter. The frequency of the generator is not important — in this case, 150 MHz was simply convenient. If the signal generator has no dBm scale, turn R16 to midpoint and adjust the signal generator for 0 dB. Insert a 5 dB attenuator. The meter should read −5 dB. Turn to the −20 dB scale while momentarily disconnecting the signal generator, then check and adjust for zero. The meter zero pot should show more control. Reconnect the signal generator and adjust for 0 dB with R17. Insert 5 dB of attenuation and the meter should again read −5 dB. Turn to the −25 dB scale and repeat the above procedure. The meter zero pot will have quite a lot of control. Note that on the −25 dB scale the needle

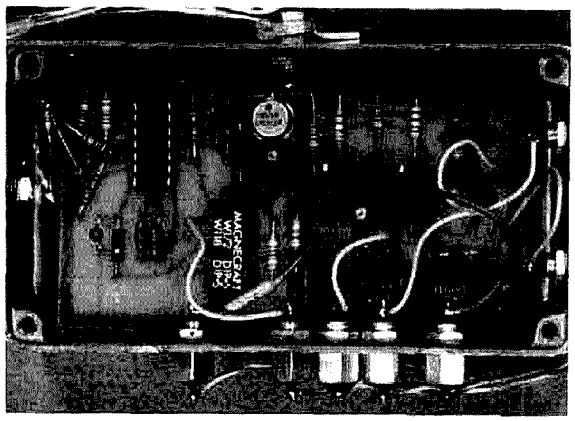


fig. 4. Parts layout in the detector-amplifier. R21, 22, and 6 are soldered between the box and the pc board. C13 is soldered on the back of the pc board under U2.

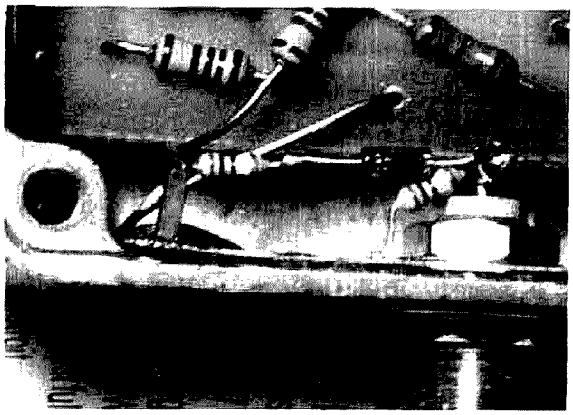


fig. 5. Close-up of the detector circuitry.

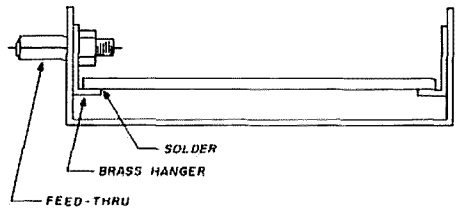


fig. 6. Printed circuit board is supported by brass shim stock.

shows some jitter or drift. This is circuit noise. This drift should be less than ± 1/10 dB at full scale. Return to the −15 dB scale, insert 5 dB of attenuation and increase output for a 0 dB reading. Turn to the −10 dB scale, remove the signal generator and adjust for zero with the right side meter zero pot. Remove the attenuator and reconnect the signal generator. Adjust R9 for 0dB. Insert attenuator; adjust signal generator

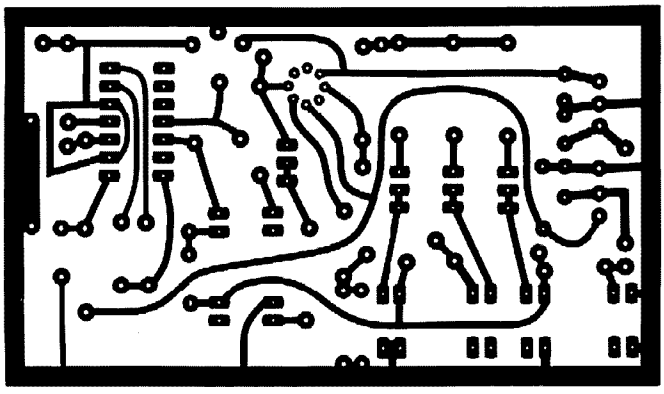


fig. 7. RF power meter artwork.

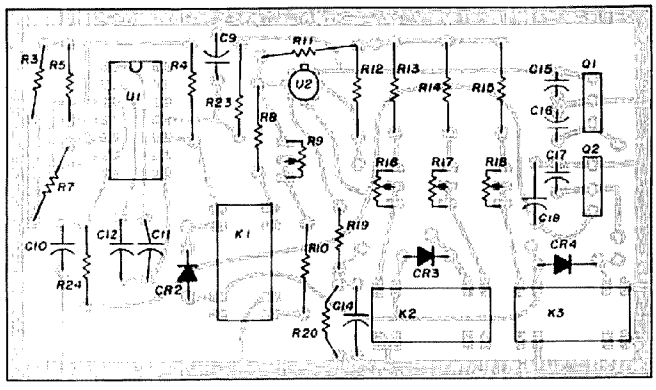


fig. 8. Component layout.

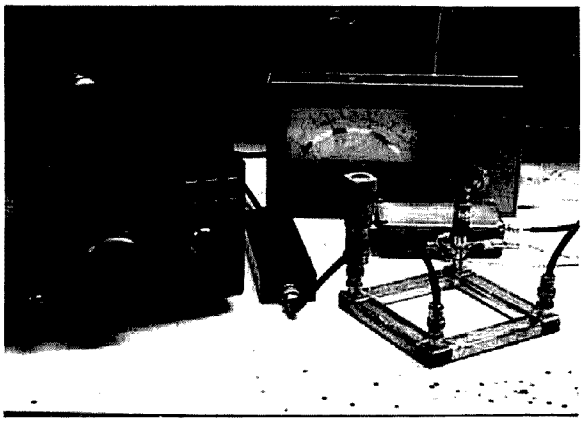


fig. 9. RF power meter is used, in a typical application, to determine hybrid coupler isolation value.

for 0dB. Turn to the −5 dB scale and remove the attenuator. The meter should read 0 dB. Insert the attenuator again and adjust the signal generator for 0 dB. Turn to the 0 dB scale and remove the attenuator. Note that the meter doesn't read 0 dB, but it should be within 1/4 dB of full scale. We are now starting to run into the nonlinear portion of the detector diode.

table 2. Parts suppliers.

C1 thru C8, C30, R1, 2, TNC connectors
Microwave Components
11216 Cape Cod
Taylor, Michigan 48180

U1, CR1
Hamilton-Avnet
Local outlets in most states. Distributor will ship COD to individuals. Cost of part approximately $25.

U2
Jameco Electronics
1355 Shoreway Road
Belmont, California 94002

K1 and Bud chassis and box
Newark Electronics
Distributors in most states.

All other parts
Digi-Key
701 Brooks Avenue, South
Thief River, Minnesota 56701

Printed circuit is available from author for $11.00. Please add $1 for shipping.

using the power meter

Figure 9 shows a typical set-up in which the power meter is used. A 70 cm hybrid coupler is checked for isolation between port 1 and port 2. The ICOM-471A provides the signal with its output reduced by a 10dB-10 watt attenuator to less than 1 watt. Further attenuation is introduced by a step attenuator.

ham radio

grounded-grid amplifier parasitics

Simple cure extends amplifier life

For several years, my kit-built amplifier with a pair of 3-500Zs had been spitting at me because of arcing at the plate tuning capacitor. I figured that either my line voltage was too high or that some flying insect was getting into the amplifier tuning capacitor and causing the arcing. This went on until the plate parasitic suppressor on the inboard tube started to smoke. (This would have been a clue for anybody who was paying attention . . . but I wasn't). I replaced the plate parasitic-suppressor and got an instant replay: I smelled burning resistor again. I didn't know what to do next, so I just lived with the stink of burning phenolic for a while. I operated the amplifier for some time, but the spitting continued. Something was wrong, but I was running out of ideas.

Nothing changed until I tried a new set of tubes and the amplifier made a noise like a shot from a 22 rifle. I pulled the plug and removed the case to inspect for damage. The problem wasn't hard to find; the grid-to-ground choke on the inboard tube had exploded its wire. One of the 200 pF capacitors from grid to ground had also exploded.

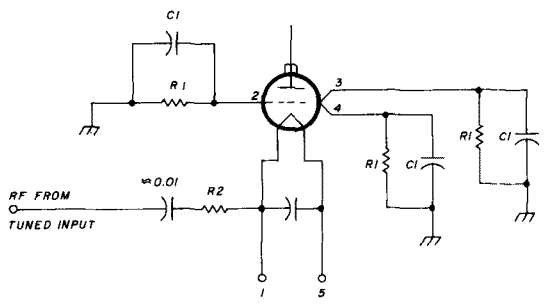
probable cause

I asked around and it seemed that other hams had experienced the same problem. The consensus was that the tube had shorted from grid to filament, causing the choke and capacitor to explode. This was confirmed by the fact that many others with this problem had found the tube to be shorted from grid to filament during the investigation following the big bang. This seemed unlikely to me because you can place a short from grid to filament on a zero-bias triode without anything cataclysmic happening. Naturally, you can't drive the cathode because the cathode is grounded, but there wouldn't be any fireworks. The answer had to be some condition that would create a grid voltage of over 2000 volts (it would take that much to destroy the 200 pF mica capacitor from the grid to ground) and more than 3 amperes of grid current which would be necesary to explode the choke wire. It had to be caused by parasitic oscillation. Light loading causes high grid current and no loading causes potentially destructive grid current and voltage. A high impedance path by the plate tank inductor would account for the "light loading" condition to VHF energy.

casualty list

Parasitic oscillation can destroy the following amplifier parts: tubes, due to grid to cathode shorts; grid current meters and shunts; zener bias diodes in the cathode circuit; contacts on the plate circuit bandswitch — usually on the 160 or 80 meter plate tuning capacitor padder contact; small chokes and capacitors in the grid to ground circuit; and, almost unbelieveably, filament transformers, because of voltage breakdown. This voltage surge is the result of the positive high voltage temporarily going to near ground potential when the tube arcs internally as the grid wires explode. With the positive high voltage at ground potential, the negative lead rises to the supply voltage, which is usually around −3000 volts in a typical amplifier. This dumps the stored energy of the HV filter capacitor into the only current path from negative HV to ground: the grid current meter and shunt resistor, which explode. This leaves an open circuit, and the negative HV arrives at the zener bias diode and the center tap of the filament transformer. Filament transformers are not usually designed to withstand high voltages, and the insulation may break down. This creates a current path inside the transformer, and

By Richard Measures, AG6K, 6455 LaCumbre Road, Somis, California 93066



R1 = 75 TO 100-OHM, 1 OR 2-WATT, NON-INDUCTIVE.
R2 = 3 10-OHM, 2 WATT, NON-INDUCTIVE, PARALLELED. FOR TWO TUBES IN PARALLEL, USE 3 10-OHM RESISTORS. FOR ONE TUBE, USE 3 20-OHM, 1-WATT RESISTORS.
C1 = 47 TO 75 pF, 500V DISC CERAMIC.

fig. 1. Grid and cathode modifications reduce tendency to oscillate at VHF.

the transformer will slowly melt unless the fuse opens. I personally know of two, a commerical pair of 3-500Z amplifiers that suffered from all of the above difficulties after the big bang.

a previous solution

Long before the 3-500Z was invented, the Collins Radio Company ran into a similar problem during the design of their 811A amplifier, the 30L-1. In order to prevent the amplifier from "taking off," a degenerative parallel R-C circuit was connected from each tube's grid-to-ground. The resistor destroyed the Q of the grid-to-ground resonant circuit. At some VHF frequency the grid structure inductance and plate to cathode capacitance resonated. This looks like a very high impedance, causing positive feedback at or near the frequency. This unavoidable situation will always develop at some frequency in any gounded-grid amplifier. But one would hope there will not be a resonant circuit in the plate compartment or in the input circuitry or associated wiring that would allow the oscillation to take place.

The Collins solution used a resistor that lowered the Q and small series capacitor that cancelled some of the built-in inductance in the grid structure, the tube base, and the socket, thereby increasing the grid resonant frequency *above* the natural resonant frequency of the grid if it had been directly grounded with a short copper strap. Other amplifier manufacturers copied the capacitor- to-ground trick, but they didn't understand that the capacitor was part of a tuned circuit. Instead, they thought it was some sort of bypass. With the belief that "bigger is better," especially when you're bypasssing, they used more than the 200 pF that Collins had wisely chosen. My troublesome amplifier used 600 pF total, with 200 pF from each of the three grid pins to ground. (See **fig. 1** for a typical circuit). I copied the Collins grid suppressor circuit. The amplifier did not oscillate. These results were published in a *ham radio* "ham note" in October, 1982.

does it work with other amplifiers?

Since that time I've learned more about the subject: whether or not an amplifier can oscillate depends on the gain of the particular tubes you have in your amplifier. New tubes may have more gain than old tubes. So the parasitic cure I used successfully in my amplifier did not always work in someone else's amplifier. After some trial and error, I found that using three separate paralleled R-C suppressors, one from each grid pin to ground, worked better in more cases than the simple circuit I described in my original article. The best values seemed to be about 50 to 70 pF for the capacitors and from 75 to 100 ohms for the resistors. This circuit worked in over 90 percent of the amplifiers.

At this point I didn't know what to do about the few remaining amplifiers that still had a tendency to take off. Fortunately, the metal-film, non-inductive resistors most people were using for the grid-to-ground resistors were acting like fast-acting fuses in the grid current path, so that no one lost any tubes. The last piece of the puzzle was furnished by a ham in Samoa who had read the original aritcle. He owned a conduction-cooled amplifier (SB-230) that used an expensive high-μ triode. He experienced a meltdown with the typical fireworks assoicated with a parasitic oscillation. The tube was ruined and the 1000 pF grid-to-ground bypass capacitor was shorted. He had installed new parts, but was worried it would happen again, since the amplifier was still spitting occasionally. I could see from the schematic that the Collins grid parasitic suppressor circuit was not going to be a possibility since there was no way to remove the existing grid bypass circuit. Any parasitic suppression would have to be done elsewhere. With a triode, this wouldn't leave you with many choices. The cathode seemed like a good bet, since EIMAC says that it takes only 27 watts to drive an 8873 to full output, so we could afford to make the tube harder to drive. This might also keep the tube from flat-topping when driven with the average 100-watt radio. A non-inductive resistor in the cathode would cause degeneration or negative feedback. This trick is often used in the emitters of bipolar transistors to prevent regeneration or instability. The trade-off is that the device is going to be slightly harder to drive. Only the drive requirement, not the power output, will be affected. I looked at the

EIMAC data sheet and noticed that the peak cathode current for an 8873 was about 1.8 amperes. Some quick calculations indicated that 11 ohms might be a good place to start. The power rating for the resistor is not easy to figure since the waveform is neither DC nor a simple sine-wave, but instead a pulse with a sine-wave shape and a duration of about 200 degrees. The peak power is 1.8 x 1.8 x 11 = 35.64 watts. The average power will be less than this for teletype operation and much less than that for SSB voice service because of the low duty-cycle.

As a simple rule, divide the peak power by 2 for teletype duty and by 3 or 4 for SSB. Dividing by 2 is recommended for the speech processor fans who like their audio to sound like creatures in a grade-B science fiction movie. Another consideration in selecting the wattage value is the fact that the average 2-watt metal-film resistor will dissipate 4 watts for at least 60 seconds with no ill effects.

Unfortunately, 2 watts is the largest size metal-film or metal-oxide-film resistor commonly available. More dissipation can be achieved by paralleling as many 2 watt units as needed. The 11-ohm resistor was installed at the socket of the 8873, in series from the cathode lead(s) to the wire that delivered the input RF drive. During peak drive conditions, 19.8 volts of RF negative feedback will be developed (1.8 amps x 11 ohms). 100 volts of peak RF output was available to drive the cathode. Losing 20 volts still left more than enough drive to give full ouput. The circuit worked. The man in Samoa was happy. No more unwelcome surprises when using the amplifier!

The same fix was tried on the 3-500Z amplifiers that had proved so difficult to tame. The same resistor bank can be used with an 8877 — in the cathode lead, of course. The cathode resistor stopped the tendency of these unruly amplifiers to oscillate. In amplifiers with a pair of 3-500Zs the peak cathode current is close to 3 amperes. This means that you can get more negative feedback voltage with fewer ohms in the cathode circuit. It was found that approximately 3 ohms of resistance would do the job. Three 10-ohm, 2-watt metal-film resistors in parallel with some space between them will work fine. These resistors were installed between the RF drive coupling capacitor — usually a 0.01 μF 1000 volt unit — and the place on the filament lead (cathode in a 3-500Z) where the coupling capacitor was originally soldered.

When 3.33 ohms are inserted into the cathode lead, the driving impedance of the cathode will be increased. This will affect the input SWR of the amplifier. This effect is greater on the higher frequency bands because the input capacity of the tubes becomes a large part of the output capacity of the tuned Pi network, and this capacity is connected through a 3.33-ohm resistor. If you don't want to adjust the tuned inputs for the 21 and 28 MHz bands, you can use a plate parasitic suppressor, made from a 47-Ohm, 2-watt resistor and four turns of Number 16 wire wound on the resistor, in place of the RF negative feedback resistor. The parasitic suppressor will not improve the linearity of the amplifier, like the RF negative feedback resistor, but it will reduce the VHF gain of the circuit and improve stability.

The 4-1000A is a stable, grounded-grid amplifier tube with plate voltages up to about 3500 volts. Above 4000 volts, the gain of the tube increases, and parasitic oscillation is possible. The quality of the amplifier tube and the frequencies of the VHF resonances are determining factors in sustaining a parasitic oscillation. A 4-1000A amplifier that proved to be unstable above 4000 plate volts was stabalized in a manner similar to the method used on the stubborn 3-500Z amplifiers. The copper, grid-grounding straps were removed. Each of the three grid pins was connected to ground through a 75-ohm resistor in parallel with a 56 pF capacitor. The screen and control grid pins were left connected to the copper plate that was previously used to bond the grids pins together. A 3.33-ohm, 6-watt resistor made from three 10-ohm, 2-watt, metal film resistors in parallel, was connected in series with the 0.01 μF capacitor that couples the drive from the tuned input circuit to the cathode (filament). After modification, the amplifier showed no sign of instability with a plate voltage in excess of 9000 volts. At this plate voltage the amplifier exhibited 15.5 dB gain. This was done to test the stability of the amplifier. Everyday use at this plate voltage is probably not going to result in normal tube life. Plate voltages this high can also produce soft X-rays, which may cause injury to the operator as well.

grid-driven amplifiers

Another use for RF negative-feedback cathode resistors is in grid-driven amplifiers, so often plagued with high-intermodulation-distortion products, or splatter. For example, a friend who owned an NCL-2000 was concerned about the interference he was causing. The root of the problem was with the 8122 tubes themselves, since the best you can expect is about −30 dB of distortion products. This is roughly 10 times worse than what you can expect from the 3-500Z. The 4CX250 series tubes have approximately the same distortion specs as the 8122. To make matters even worse, the NCL-2000 design allowed grid current to flow freely during modulation. This causes the tube's plate current curve to take a nasty jag at low values of plate voltage. My friend was able to make the amplifier acceptably clean by installing three 2-watt, 51-ohm, metal-oxide-film resistors per tube, with one resistor in series with each of the three cathode connections per socket. He also changed the tap

on the 50-ohm grid-swamping resistor to avoid driving the control grid into conduction. This depends on how much driver power you have. The result was appreciated by all concerned.

There are other external-anode triodes besides the 8873, 8874, and 8875s that are capable of taking off at VHF or even UHF frequencies. I have recently talked to two people who experienced instability problems with the 8877. For some reason, the problem seems to occur only when these tubes are used in HF amplifier circuits. Perhaps this is because of the extra lead-lengths required in a HF amplifier design and by the fact that these tubes have excellent gain up into the UHF region. One of the 8877 amplifiers was a DTR-2000. The owner had discovered one of these no-longer-made amplifiers in an unopened box in a dealer's warehouse. He bought it for not much more than the price of a new tube. He was delighted with his "find" until the "big bang" occurred during his first day's use.

With a 3-500Z parasitic, the grid may weld to the cathode. I've occasionally seen grid wires blown loose and rattling around inside the bottom of the glass envelope. Such a tube may continue to work. The 8877 in the DTR-2000 also had a wire from the grid blown loose, but from only one end. The free end of the wire had shorted to the plate of the tube.

Another problem with the DTR-2000 is that 5.9 volts is applied to the filament, which, according to EIMAC, should never have more than 5.25 volts. This situation will appreciably shorten the life of the 8877's oxide-coated cathode. Some owners have corrected the problem by installing a 0.1 ohm, 10 watt resistor in series with the filament.

The important thing to keep in mind is that the average amplifier tube, with average gain at VHF, probably will not oscillate in a typical HF amplifier design. The problem shows up when you happen to get a unusually *good* tube — with lots of VHF amplifying ability — in a HF amplifier circuit.

the why of it

If you wanted to build an oscillator, you would need at least one tuned cirucit, an amplifier, and a feedback path. If you had two tuned circuits, one for the input and one for the output, your chances of building a successful oscillator would be even better. Keeping these facts in mind, I started sniffing around the input and output circuitry of my amplifier with a dip-meter. The bandswitch was set to 40 meters — the same band in which the big bang was heard. The drive coupling capacitor at the cathode of the 3-500Zs, which connects to a short length of 50-ohm coax, showed a good dip at 110 MHz. The lead from the plate of the tubes to the plate tuning capacitor showed a good dip at 105 MHz despite the presence of the parasitic suppressor. The plate circuit dip could be moved to 110 MHz by slightly adjusting the plate tuning capacitor. The grid showed a dip near 90 MHz. We have two tuned circuits. We have a feedback path through the "grounded" grid. The 3-500Zs are rated at 110 MHz. It should be capable of sustaining oscillation. It does.

These resonances are nobody's fault. They are caused by the laws of physics and they cannot be eliminated by any practical amplifier design. The way to control the problem is to use non-inductive resistors in the input and output circuits to destroy the *Q* of the VHF resonances. Most people are accustomed to using a resistor in the plate lead to control parasitics, but the idea of using such a device in the cathode lead is sadly missing in most HF amplifier designs. This is sad because the cathode lead is an ideal place to accomplish the job, since a resistance in the cathode lead will cause desireable negative feedback — which the plate circuit cannot do. If I had to pick just one place to put a parasitic suppressor, the cathode would be a good choice.

If you're thinking that your amplifier is immune to the problem, uou may be right — for the particular set of tubes that are in service . . . . The next time you're doing your annual spring cleaning inside the amplifier, check the plate lead for resonances with a dip-meter; you're going to get a nasty surprise. The drive coupling capacitor will also show a resonance whose frequency is mainly dependent on the length of coax that connects to the input bandswitch. The schematic does not show any VHF tuned circuits, but they're always there. Remember this when you plug in that hot new set of tubes. Two-watt resistors are cheaper than new tubes.

If you own an amplifier that sometimes snaps or spits, this isn't something to ignore unless you enjoy fixing broken linear amplifiers. The amplifier is trying to tell you something — if you're paying attention, you *can save yourself some expensive grief.*

neutralizing grounded-grid amplifiers

Why can't a grounded-grid amplifier be neutralized, like a Class AB, grid-driven amplifier? In EIMAC's book *Care and Feeding of Power Grid Tubes*, it's stated that grounded-grid amplifiers don't normally need to be neutralized. This is not a very comforting statement, considering that there appears to be no way to neutralize a single-ended, grounded-grid amplifier, even if you want to. Gonset tried to neutralize a four-tube 811A amplifier with notoriously poor results. If you own one of these, the Collins circuit will cure the problem.

references

1. Richard Measures, AG6K, "3-500Z Tube Failure," *ham radio*, October, 1982, page 78.

ham radio

universal oscillator circuit

Test crystals over a 200:1 frequency range

For more years than I care to remember, I've been collecting crystal oscillator circuits with the hope that one day I'd stumble across the ultimate oscillator circuit. The ultimate circuit would allow me to test the oscillating frequency of all types of crystals from 100 kHz to 20 MHz. No tuning or parts changes would be needed; I'd just swap the crystal and watch the activity on some kind of meter. I'd also be able to measure the frequency as accurately as possible.

Over the years I've yet to see a circuit with this capability that could be duplicated without too much trouble and, most of all, some kind of explanation of why things were done as they were including all relevant technical details, complete with accurate parts information so you'd know what can and what cannot be substituted. Being in radio repair myself, I felt it would be very nice to have something to count on; not finding anything really suitable, I finally had to come up with some ideas myself. The circuit shown in **fig. 1**, which I call the OmniTek oscillator, shows the results. It may not be the ultimate oscillator circuit, but so far I've not seen one more versatile or better suited for my needs.

circuit description

Figure 1 shows the oscillator, Q1, and its associated parts: Q2 (the buffer) and Q3 (the emitter follower with dual output, one for the indicating meter and the other for a counter or other uses). I've seen variations of it before, but not with the 200:1 range this one has. The secret seems to be in the 10 mH choke (scramble-wound miniature coil on a ferrite core) on the drain of Q1. Having tried all kinds of combinations, including other types and makes of L1, I found that only the specified choke worked well and consistently every time.

Various versions of this oscillator — including a handheld test unit with meter and also one that replaced a master multiple crystal oscillator that used a tube (in that well known 6BH6 circuit) — were built.

The Activator button is for low activity or 3rd overtone crystals that may need an additional jolt to start. Most of the time, however, it isn't necessary and could certainly be omitted. If you don't want to, or cannot, calibrate the oscillator for a 32 pF load by plugging in a known crystal calibrated for 32 pF, omit capacitors C1 and C4. Just be sure that the values of C2 and C3 are correct because their ratio, 51 to 56 pF, is very important for the correct operation of the oscillator. They're also the correct values for a very close approximation of a 32 pF load. Use 5 percent silver micas or NPO ceramics here. As a matter of fact, all capacitors in the vicinity of Q1 — that is, C2, C3, C4, C5, C7 and C13 — should be either silver micas or NPO ceramics. (I prefer the ceramics because they're so much smaller.) The trimmer, C1, if used, is often an N450 or N750, though the temperature coefficient really doesn't matter much. All other capacitors may be standard, and will not affect the operation of the oscillator at all. All resistors should be at least 1/4 watt, except the pot, which should be 1/2 watt. The meter can be just about anything you can find, but full scale deflection sensitivity should not be much over 200 μA; if it is, you won't get a good (i.e., more than half-scale) indication on the meter on the higher frequencies. The diode across the meter is used to limit the maximum voltage on the meter (to about 300 mV) because on crystals below 10 MHz the output may be high enough to damage the more sensitive meter movements. The

By Robert H. Fransen, VE6RF, 227 Cottonwood Avenue, Sherwood Park, Alberta, Canada T8A 1Y3

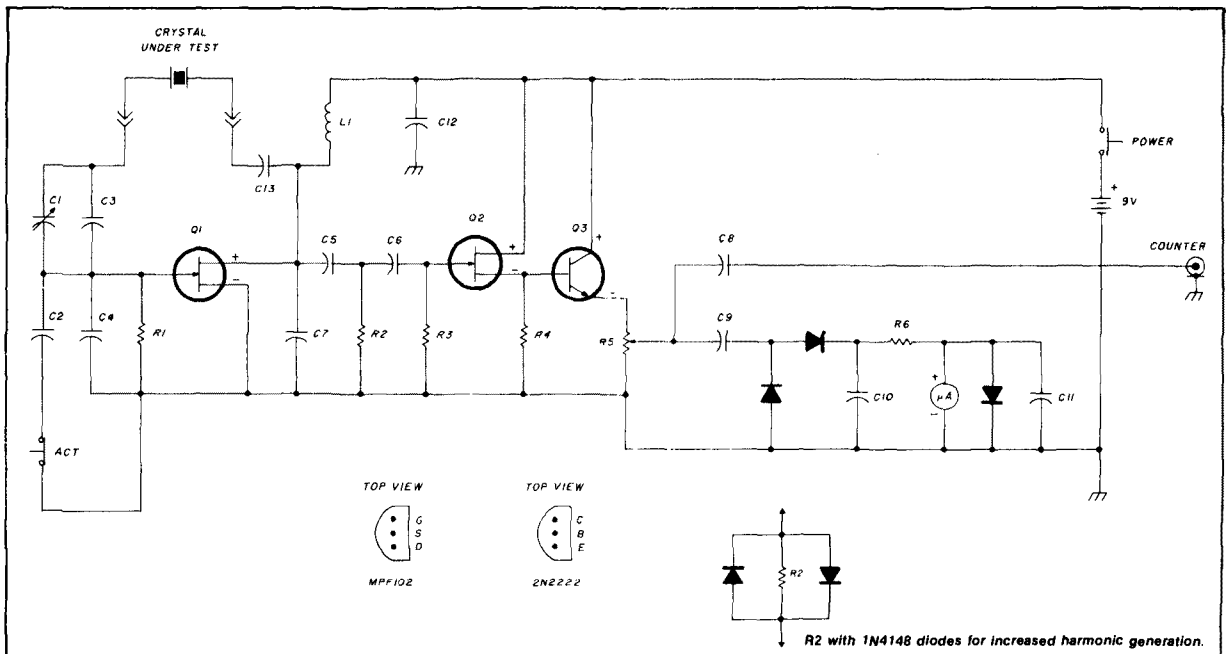


fig. 1. 100 kHz to 20 MHz crystal oscillator (32 pF load).

counter output is tied to the wiper of the pot but could also be connected directly to the emitter of Q3, if desired.

To check the battery (with the terminals mounted downward in the case of the portable version), I drilled a small hole through the bottom below the + terminal. By sticking a probe through the hole (don't short the probe to the case) you can test the battery voltage without taking the box apart.

application

Although the unit is intended for checking crystals, it can also be used as a rough-and-ready signal generator or spotting calibrator. If you use it for spotting and you want plenty of harmonics, connect two diodes (1N4148) in parallel back-to-back across R2 (see insert in **fig. 1**). The only effect on the oscillator characteristics will be an increase in the harmonic output.

If you have a circuit board with soldered-on crystals, checking the crystals can be very difficult because taking them off the board very often destroys them. A much easier way is to cut the traces to the crystal and put two No. 18 sewing machine needles (mounting shaft ground off) in the oscillator socket holes. Press them against the traces of the board; the oscillator will indicate the crystal quality.

If you don't intend to use the circuit for this purpose, C13 can be omitted. It's there only to keep DC off the crystal socket. Omitting C13 allows you to test the battery voltage on the socket connected to the drain of Q1 so no holes have to be drilled. Never try to insert a crystal without holding the box in your hand; if you do, static will damage Q1. A good indication of damage to Q1 (gate leakage) is if the unit oscillates only on the higher frequencies. Although the PWR button can also be a switch, whenever you

Crystal oscillator parts list.

C1	2.5 – 11 pF
C2	51 pF
C3	56 pF
C4	3 pF
C5	560 pF
C6	100 pF
C7	270 pF
C8	100 pF
C9,C10,C11	0.01 μF
C12	0.47 μF
C13	1500 pF
R1	2.2M ohms
R2	2.2K ohms
R3	1M ohms
R4	4.7K ohms
R5	500 ohms
R6	150 ohms
Q1	MPF 102
Q2	MPF 102
Q3	2N2222

All diodes: 1N270
L1 is Hammond Number 1530 C102
10 μH resistance is 1000 ohms (@ 38 mA, maximum).
Meter is 140 μA at 140 microvolts F.S.

Minimum current at 100 kHz is about 6 mA. Maximum current at 20 MHz is between 14 and 22 mA, depending on the gain of Q1. Frequency shift over a supply voltage range from 5 to 10 VDC is less than 0.5 PPM. Battery is 9 VDC type, Number 1604. Some waveform distortion takes place below about 3 MHz. The ACTivate button is used to test and start 3rd overtone crystals that may need more feedback to start. The 5th, 7th, and 9th overtone crystals will probably not oscillate in this untuned circuit. Basic circuit (with C1 and C4 left out) is for a 32 pF load. Meter can be up to 200 μA full scale, though 50 to 100 μA are preferred. Minimum and maximum currents are for the whole circuit.
R2 with 1N4148 diodes for increased harmonic generation.

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check the battery make sure there's no crystal installed so the circuit uses maximum current.

Construction and wiring are not critical. Just try to keep the leads near Q1 as short as possible and install Q1 in a manner that will allow it to be replaced easily, because it's easily destroyed.

If you want to keep things simple and not use C1 or C4 or the ACT button, install C2 (51pF) and C3 (56pF) and check your frequency. If it's too high, a gimmick wire capacitor across C2 will bring the frequency down. If the frequency is too low, replace C2 with a 47 pF capacitor and try the gimmick capacitor again on C2 — that is, if you have a good calibrated crystal. If you don't, install the caps and forget about calibration. Also keep in mind that 3rd overtone crystals don't oscillate at precisely 1/3 of their frequency because they're series-calibrated. It's understood, of course, that the indication on the meter is strictly relative. But after a bit of use, and a knob on the pot with a calibrated skirt, you'll get the hang of it pretty quickly and know what to expect.

Other variations of this circuit are possible. Since supply voltage and load variation don't, for all practical purposes, affect the frequency (keep your hands away from the crystal), further experimenting may be in order, perhaps with other 10 mH chokes.

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ham radio TECHNIQUES

Bill Orr W6SAI

surface-wave OTH radar — more QRM?

The Wireless Institute of Australia is reportedly developing an experimental over-the-horizon (OTH) radar that operates only over the sea. The radar transmits a vertically-polarized radio wave close to the sea surface, inducing electrical currents in the water. This causes the radio wave to adhere to the sea surface and therefore travel around the curvature of the earth. There is a possibility that the reflected energy will couple with the sea surface for the return journey.[1]

When I read this, it rang a bell; somewhere I'd heard that experiments have been run in the Caribbean in which long-distance VHF communication was established between ships by placing vertical Yagi antennas very close to the waterline. When the antennas were raised more than a few feet above the sea, the signals dropped in strength. The guess was that the layer of ocean moisture directly above the surface of the water provided the conduit for the radio wave.

That's all I know about the Australian report and the Caribbean experiment. If any reader knows more about the ocean-wave experiments, I'd certainly like to hear about it. How about a test between California and Hawaii?

the "underwater antenna"

Jokes and tall stories about underground transmitting antennas and antennas immersed in water have appeared in Amateur literature for decades. I've also heard that an antenna immersed in water will not only work, but because of the dielectric constant of the liquid, be markedly smaller in size, for a given radio wavelength, than antennas not immersed in water. Sounds like a great idea — a 160-meter antenna in the back yard swimming pool!

In 1978 a British patent (GB2,001,804) was filed by the Plessey Company for an "underwater antenna" (**fig. 1**). The idea proposed in this patent is a variation of the principle of dielectric loading. According to an article in *Radio Communication*, the patent application reads, in part, as follows:

It has been proposed to submerge an antenna consisting, for example, of a metallic rod in water, but this has been found to suffer from the practical disadvantage that through contamination and absorption of carbon dioxide into the water, degradation results and the antenna efficiency rapidly deteriorates.

It is therefore proposed to use an antenna structure surrounded by water or similar acceptable liquid by including a sealed container shaped so that the element is completely surrounded by the liquid. This can take the form of a sealed glass or plastic container filled with water (some anti-freeze can be added for low temperature conditions). The process of filling and sealing the container is preferably carried out under chemically-clean conditions.[2]

The patent claimed that an antenna rod length of 15 cm (about 6 inches) used at 100 MHz gave an increase in signal strength of over 200 percent compared with a rod of the same length in free air.

Discussing this antenna, Pat Hawker, G3VA, says, "The idea of surrounding an element with water reminds me of a problem known to exist with some wideband television receiving arrays: a significant fall-off in performance on the higher frequency channels when it rains and water collects on the elements. Clearly what is happening is that the resonant frequency of the array is being lowered by the rain — further proof of the effects of dielectric loading. . . . But one foresees an unhappy operator reporting: 'Sorry OM, signals are fading, my antenna has sprung a leak.'"

great circle maps

Not easy to find, these days. I wanted a large Great Circle Map centered on San Francisco. But where to get it? I did a little footwork and found out that these maps can be obtained from the Office of Distribution Services of the Defense Mapping Agency (Hydrographic Center). The mailing address is: DMA-ODS, attention DCCP, Washington, DC 20315. Great Circle maps cost $5.50 each and may be ordered by stock number from the catalog. To order the catalog, send $2.25 to the address above and request catalog No. CAT-P2V10 which, according to the obliging individual who answered my call, is a "goldmine of information."

more on the 160-meter end-fed antenna

In my last column I mentioned my long, 160-meter end-fed antenna, series tuned with a capacitor and matched to 50 ohms with a shunt coil. I've had it on the air for some weeks now and find it to be the best antenna that I've been able to put up on this particular piece of property, considering the zoning restrictions. Best DX to date has been Japan and Siberia, zone 19.

For those who have less space, the quarter-wave Marconi is still a good

antenna. It can be easily matched to a 50-ohm feed system by the technique shown in **fig. 2**. The antenna is cut to your favorite operating frequency in the band by the formula: length = 234/f (MHz). For 1825 kHz, the antenna is about 128 feet, 3 inches (39.09 meters) long. If the antenna is entirely vertical (an unlikely assumption), the feedpoint resistance (R) at resonance will be about 36 ohms. As more and more of the antenna lies in the horizontal plane, the feedpoint resistance decreases. In my tests, with most of the antenna wire running horizontally about 40 feet above ground, the feedpoint resistance ran close to 15 ohms.

You see that a simple L-network (A) can be used, made up of a series- connected coil and a shunt capacitor. The coil is quite small, but the capacitor value is rather large. The coil can be a small B & W "Miniductor" about 2 inches (5.08 cm) in diameter, with a tapping clip for adjustment. Only 2.5 μH is required to do the job under all circumstances (see chart). Note that maximum inductance is required when the feedpoint resistance is one-half the value of the input resistance of the network.

The tapped coil presents no problem, but obtaining the shunt capacitor can be vexing. Most end-fed 160-meter Marconi antennas fall into a feedpoint resistance range of 10 to 25 ohms. This calls for a shunt capacitor value of approximately 3500 to 1500 pF. The total capacitance can be made up of several "postage stamp" silver mica capacitors placed in parallel with a large variable capacitor. In my case, I have a 900 pF variable capacitor picked up at a flea market and a rotary switch that adds fixed capacitance at 500 pF per switch position.

This combination allows excellent antenna matching to be obtained all across the 160-meter band. I use an SWR meter to determine antenna match (the meter being placed between the network and the short coax line to the transmitter). A practice run, tuning up every 25 kHz across the band, provides logging points for the coil and capacitor settings so that no time is lost when I want to QSY from 1810 kHz to work a UA0 calling CQ on 1915 kHz.

The chart also shows why it's sometimes difficult to get a good match to a low frequency mobile antenna. The matching coil becomes quite small for

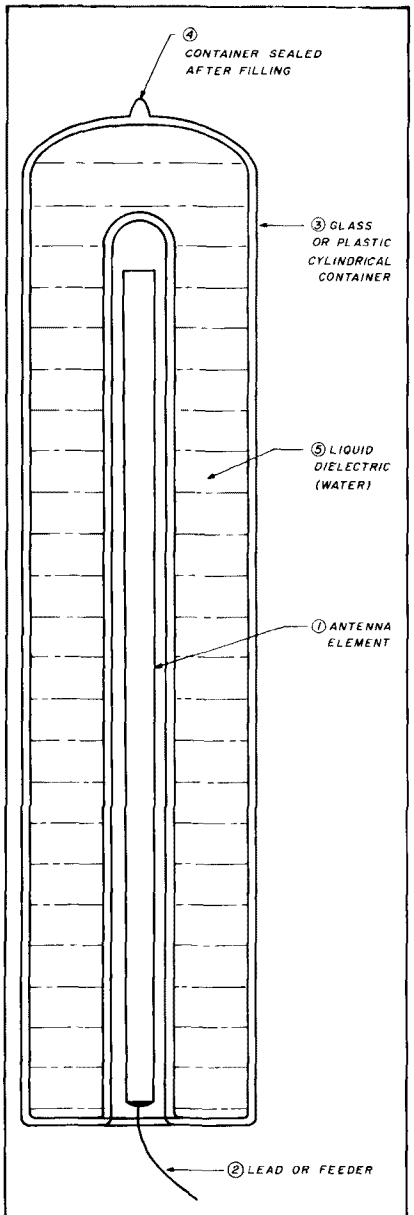


fig. 1. Short "water-jacketed" dielectric-loaded antenna as disclosed in UK Patent Application GB 2,001 804 by the Plessey Company.

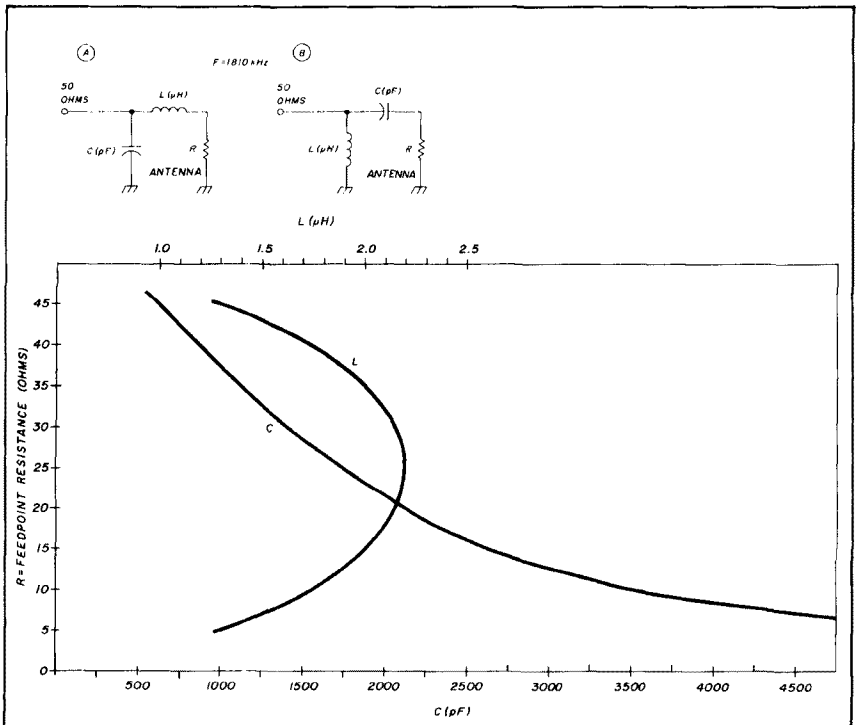


fig. 2. L-network components for 160 meters and chart to determine component values for network A.

low values of feedpoint resistance and the shunt capacitor becomes quite large!

match for the HF mobile antenna

The 80- or 160-meter mobile antenna presents a matching and loading problem. It's generally agreed that center loading provides the greatest operating efficiency for such an antenna, and many Amateurs have had success with an 8-foot (2.43 meter) antenna loaded in this fashion. Unfortunately, the feedpoint impedance of such a loaded antenna on the 80 and 160-meter bands, runs in the region of 20 ohms, of which only about 0.5 ohms is radiation resistance, the balance being made up of loading coil losses.

The B-network shown in **fig. 2** is often used for mobile whip antennas. All that's required is a small shunt inductance in the range of 1.5 to 2.5 μH for 160-meter operation. The series capacitance can be the actual antenna adjusted to provide a capacitive reactance at the base, that is, one that's slightly shorter than its resonant length.

Adjusting the antenna is quite simple. With the base matching inductor removed, a two-turn coil is connected between the base of the antenna and the grounding point on the vehicle directly below the antenna. A dip oscillator is used to set the antenna on frequency. Loading coil turns are adjusted to provide indication of antenna resonance. The base coil is now inserted in place of the dip oscillator loop and an SWR meter is placed in the coax line to the transceiver. Reduced power is applied to the antenna at the resonant frequency and the antenna is readjusted to resonance by pruning the loading coil. Lastly, the base inductor is adjusted for lowest SWR at the antenna resonant frequency.

The adjustments are slightly interactive and the presence of the experimenter in the immediate field of the antenna will tend to detune it a bit. The process sounds tricky, but it really isn't . . . it just takes a bit of patience and common sense.

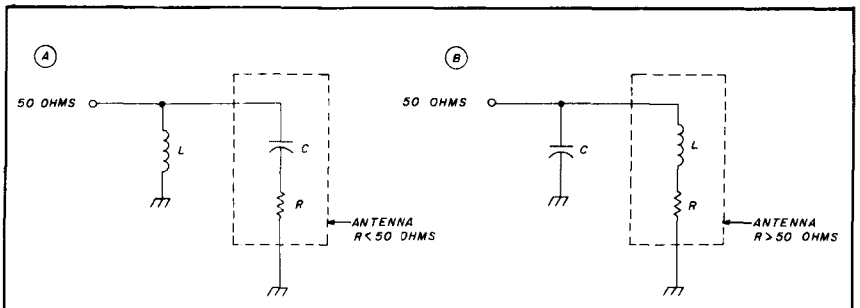


fig. 3. Antenna has reactive feedpoint impedance if not resonant at required frequency. (*A*) Capacitive reactance if shorter and (*B*) inductive reactance if longer.

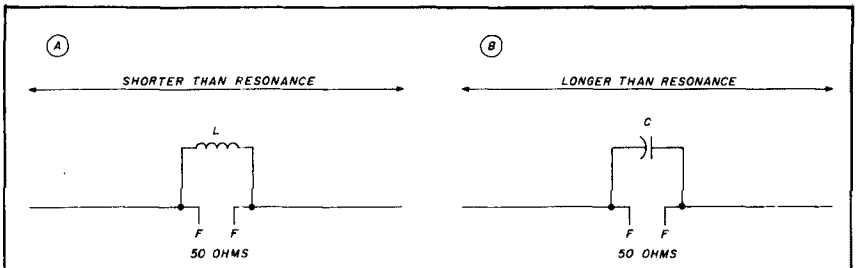


fig. 4. Yagi driven elements shorter (*A*) and longer (*B*) than resonance can be matched to coax line with appropriate type shunt reactance.

the simplified L-network

The two matching networks shown in **fig. 2** can be redrawn as shown in **fig. 3** in which the series component is represented by an off-resonant antenna. **Figure 3A** illustrates the case in which the antenna is shorter than the resonant length. **Figure 3B** shows the case in which the antenna is longer than resonant length. The circuit shown in **fig. 3A** is used in some Yagi beam antennas, where the inductor takes the form of a coil, or hairpin, placed across the feedpoint in shunt with the driven element. In this case, the driven element is shortened slightly to provide a capacitive reactance at the feedpoint.

By using the reactance of the antenna element as one arm of the L-network, either by lengthening or shortening the element past the resonant point, an effective and inexpensive matching system that requires only one additional shunt element — either a capacitor or an inductor — can be made.

Use of this matching scheme with a balanced Yagi element is illustrated in **fig. 4**. In **fig. 4A** the driven element is made slightly shorter than resonance and an inductor is placed across the feedpoint. The inductor may take the form of a balancing device so that impedance transformation and transformation to a coaxial line is accomplished with the same device.[3] Most high frequency commercial matching devices take this form because it's easier to make a waterproof inductor that will withstand high power than a suitable capacitor. (Matching systems of this general type are discussed in detail in the new edition of the *Beam Antenna Handbook*.*)

the EME directory

A reprint of the WA1JXN 144-MHz EME (Moonbounce) directory is now available. Listing EME operators worldwide, including their addresses and the equipment they use, this 16-page com-

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references

1. B. Martin, "The Woodpecker Project," *Amateur Radio* (Australia), August, 1985, page 13.
2. Pat Hawker, G3VA, "Technical Topics," *Radio Communication*, (England), September, 1985, page 708.
3. Bill Orr, W6SAI, and Stu Cowan, W2LX, *Beam Antenna Handbook*, pages 164-175.

ham radio

short circuits

reflector antennas

Eqn. 1 in W1JR's February column ("Reflector Antennas, Part 1," page 54) should be corrected to read as follows:

$$G = 10 \log [0.55 \cdot 4\pi A/\lambda^2]$$
$$= 10 \log (6.9A/\lambda^2)$$

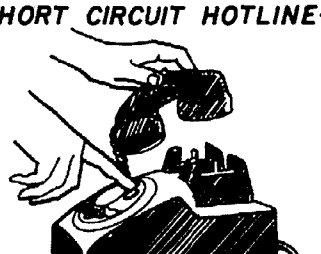
calibrated S-meter

The value of the Pin 12 resistor shown in **fig. 3** of W7SX's article, "A Calibrated S-Meter" (January, page 23) is 2000 ohms.

upside-down battery

In **fig. 8** of YB9ATA's February article, "Two-Tone Signal Generator," the battery was inadvertently shown upside-down.

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(See "Publisher's Log," April, 1984, page 6, for details.)

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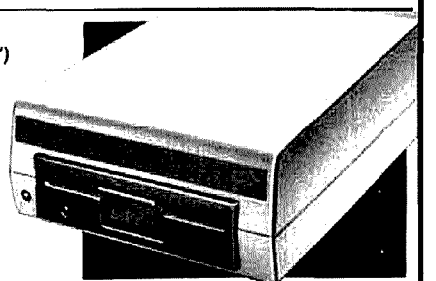
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computer control of ICOM R-71, 271, 471, and 751 radios

Extend performance and versatility by combining analog and digital techniques

Does the idea of using a computer to control your radios conjure up images of driving elaborate remote bases with touchtone commands from a handheld? It needn't. There are less exotic uses of computer control that can really make things easier around the shack.

What else can you do with computer control? Suppose you want to monitor a net, a beacon, or a bulletin, but can't be near the radio. No problem — *get a computer.* It can turn on the radio at a prescribed time (say 18 minutes past the hour), tune it (perhaps to WWV), turn on a tape recorder (to record a minute's worth of propagation bulletin), and then turn everything off. You can listen to the tape at your convenience.

Working satellites such as OSCAR-10 is another area in which a computer can generally simplify operation. If you've ever listened to an Amateur satellite, you've found the passband filled with "aaahhh" or machine-gun strings of dits. Why?

To use a satellite, you must transmit on one band and receive on another; few Amateurs are adept at operating two radios simultaneously. To complicate matters, the radios are on opposite sidebands . . . to tune, you turn one knob clockwise and the other counter-clockwise — a trick easily learned by 5-year-olds, but not by adults. Doppler shift also has to be accounted for. As a result, Amateurs spend half their time trying to find their signal in the satellite's passband. The solution? *Get a computer.* Tune the receiver and let the computer read its frequency and tune the transmitter. A piece of cake!

Interested in these and other applications of computer-controlled radios? Computer control of ICOM's latest series of radios can be an interesting project, and requires only some simple hardware. Read on.

computer-controllable radios

ICOM has been manufacturing computer-controllable radios for many years, beginning with the 701/211/245 series and continuing with the R-70/720/251/255/260/451 series. Control of the earlier radios was generally limited to changing frequency and mode. ICOM's latest series of radios, the R-71, 751, 271, and 471, allow additional radio functions to be controlled by the computer, including the 32 memories. The newly announced 1271 will almost certainly be controllable in the same way. Like earlier ICOM radios, the interface uses a parallel handshake, with all radios daisy-chained on a common bus. Unlike earlier radios, the computer interface isn't included with the radio, but must be purchased separately. The interface is the EX-309 Microprocessor Interface Connector, which sells for $37 and consists of a small board, approximately 2 inches by 2 1/4 inches (5.08 by 5.72 cm), containing two octal latch ICs, a voltage regulator, and a 24-pin female IEEE-488 (Centronix-type) connector. (Previous ICOM radios used pins on the 24-pin Molex accessory connector for computer interface signals.)

required interface board

The EX-309 interface allows external 8-bit data to be gated onto and off of the radio CPU's internal 8-bit data bus. Signals available on the external connector are an 8-bit bidirectional data bus, a service request ($\overline{SRQ}$) line, read (RP) and write (WP) request lines, a data valid ($\overline{DAV}$) line, and squelch and send lines that parallel signals on the Molex connector (see **fig. 1**). Ground and 13.8 volts (100 mA maximum) are also available on the connector. You'll need 14 TTL lines on your computer: eight bidirectional, four output, and two input. If you don't already have these lines, you can add them using a parallel interface adapter chip (e.g., a 6522, 6820, 6821, or 8255) or build them out of TTL latches.

The EX-309 is easy to install once you realize that

By Richard Bisbey II, NG6Q, Suite 1001, 4676 Admiralty Way, Marina del Rey, California 90292-6695

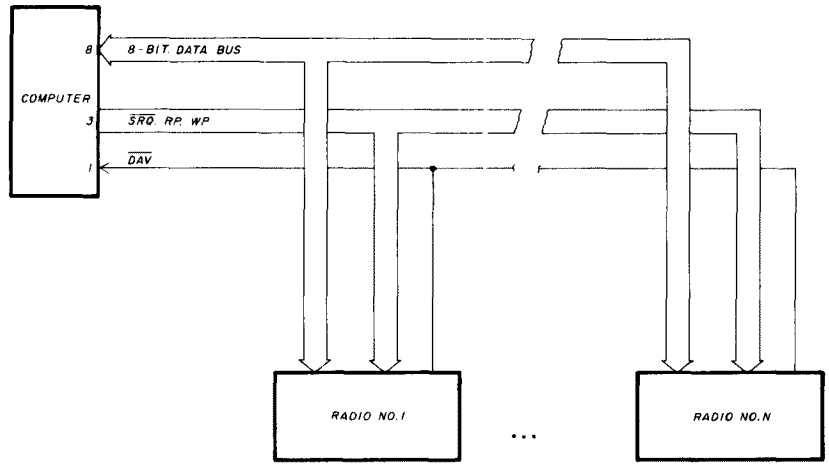


fig. 1. Radio to computer connection.

you can get the connector through its mounting hole if you insert it "end first." (You'll have to remove either the metal plate or the rubber dust cap covering the mounting hole first, of course.) The board mounts on the rear left bottom of the 271/471 and on the rear right side of the 751. There are RF chokes on the EX-309 board. However, you may want to insert ferrite beads in the lines to minimize external signals entering the radio and being re-radiated between the connector and the board. If you go to the trouble of desoldering the connector, you might consider replacing it with a DB-25, which takes the same space, is cheaper, and is more readily available. You don't have to worry about maintaining compatibility with other accessories, since the only ICOM accessory that uses the EX-309 is the CT-10 RTTY TU, which is not currently being imported into the United States by ICOM America. If you stick with the original connector, you can get its ribbon connector male mate for $7.95 from Jameco, 1355 Shoreway Road, Belmont, California. Be sure to specify the spring type, although the screw type will work satisfactorily.

The EX-309 has three connections inside the radio: data bus (P4), control bus (P5), and send/squelch (J3). For the 271/471, P4 goes to Logic Board J3, P5 to Logic Board J1. For the 751, P4 goes to Logic Board J15, P5 to Logic Board J10. If you are installing the EX-309 in a 471, be sure to correct your schematic by adding the 13.8-volt line to Pin 9 on P5. The third connection, from J3 to the send/squelch lines, is made to Front Panel P12 on the 271/471, to the AF VR board on the 751. *This connection is not listed in the instruction sheet that comes with the EX-309.* If you have more than one radio on the external bus, you probably won't want to make this third connection. If you were to make the third connection, the squelch and send lines for all the radios on the bus would be connected in parallel, and you couldn't remotely key individual radios or tell which radios were or were not squelched. Also, the squelch line on the 751 is 8 volts and requires a 5.1 volt zener diode to ground at the connector to make it TTL-compatible.

Once installed, the interface is easy to use. The protocol to use in communicating with the radio is as follows:

1. Drop $\overline{\text{SRQ}}$ to 0V (to get the radio's attention).
2. Use procedure A or B (see below) to send or receive a byte.
3. If not finished, go back to step 2.
4. Raise $\overline{\text{SRQ}}$ to 5V (to tell the radio you are finished).

To send a byte to the radio, follow *Procedure A*, as follows:

1. Set 8 bits of data on the data bus.
2. Raise WP to 5V (to tell the radio you are writing data to it).
3. Wait for the radio to drop $\overline{\text{DAV}}$ to 0V (to ACK receiving the data).
4. Drop WP back to 0V (to ACK the ACK).
5. Wait for the radio to raise $\overline{\text{DAV}}$ to 5V.

To receive a byte from the radio, follow *Procedure B*, as follows:

1. Raise RP to 5V (to tell the radio you are reading data from it).
2. Wait for the radio to drop $\overline{\text{DAV}}$ to 0V (to ACK sending the data).
3. Get 8 bits of data from the data bus.
4. Drop RP back to 0V (to ACK receiving the data).
5. Wait for the radio to raise $\overline{\text{DAV}}$ to 5V.

Only one command can be issued to the radio each time the $\overline{\text{SRQ}}$ line is lowered. Also, there is a mini-

mum time the $\overline{SRQ}$ line must remain high before it can again be lowered as well as a minimum time between lowering WP and raising $\overline{SRQ}$ when sending data to the radio. This limits the rate at which commands can be issued. Generally, the radios can accept up to about 50 commands per second. Thus, the minimum dwell time for frequency-hopping, spread-spectrum uses would be 20 msec (subject to the settling time of the radio's PLL). Finally, if a radio fails to acknowledge an RP or WP within one second, it is either not connected or not powered on, or is simply otherwise occupied (e.g., scanning).

commands and their operands

Each 8-bit byte on the data bus is actually two 4-bit nibbles. The four most significant bits of each byte encode the command (or operation code). They are:

- 1x—Band Data (read only).
- 2x—Frequency Data (read/write).
- 3x—Mode Data (read/write).
- 4x—Offset Data (read/write).
- 5x—Set Memory/VFO (write only).
- 6x—Memory Read/Write (write only).

The four least significant bits of each byte encode address and data operands. The first hex digit is the radio's address. It is always written to the radio. Valid addresses are:

- x1—HF — R-71 or 751
- x2—50 MHz
- x3—144 MHz — 271
- x4—220 MHz
- x5—440 MHz — 471
- x6—1200 MHz — 1271

(Too bad ICOM didn't leave x6 for 902 through 928 MHz and move 1200 MHz to x7.) Data operands follow the address, and, depending on the command, may be either written to or read from the radio. Each data operand consists of a string of hex digits delimited by "D" and "E." The radio will ignore all write data between the address operand and the first delimiter, "D."

specific command information

The following is a description of each command along with its operands. The actual hex bytes exchanged with the radio are shown, with the command in the high nibble and the operand values in the low nibble.

Command 1 — band data. This command allows the computer to read the frequency range of a radio. The frequency range is returned as:

1D 1m 1m 1m 1m 1m 1m 1E 1D 1n 1n 1n 1n 1n 1n 1E

where mmmmmm and nnnnnn are the upper and lower frequency limits in tens of kHz. To request the frequency range of a 471, send the hex byte 15 (command 1, address 5), and then read back the following sixteen bytes:

1D 10 14 14 19 19 19 1E 1D 10 14 13 10 10 10 1E

i.e., 0 4 4 9. 9 9 to 0 4 3 0. 0 0 MHz.

Command 2 — frequency data. This command allows the computer to read or write the radio's frequency. The frequency is a nine-digit number; the most significant digit is GHz, the least significant is tens of Hz. Attempts to set a radio to a frequency outside its band limits are ignored. To set a 271 to 145.67893 MHz, send:

23 2D 20 21 24 25 26 27 28 29 23 2E

i.e., 0 1 4 5. 6 7 8 9 3 MHz.

The radio will set unsent digits to zero, so the sequence: 21 2D 20 20 21 2E would set a 751 or R-71 to WWV at 10 MHz. To read a radio's frequency, send hex 2# (where # is the radio's address) and read back eleven bytes (2D, nine digits with 2 in the leftmost nibble, and 2E).

Command 3—mode data. This command allows the computer to read or write the radio's operating mode. Operand values are:

0 — LSB
1 — USB
2 — AM
3 — CW
4 — RTTY
5 — FM
6 — CW-Narrow
7 — RTTY-Narrow
8 — LSB
9 — USB
A — AM
B — CW-Narrow
C — RTTY-Narrow
D — FM

Not all modes are possible on all radios. For example, the 271 and 471 lack RTTY capability. Thus, 4, 7, and C wouldn't make sense for those radios, and, in fact, would leave the radio in an indeterminate mode. The sequence 31 3D 3C 3E would set a 751 to RTTY-Narrow, while the sequence 35 3D 31 3E would set a 471 to USB. To read mode data, send hex 3# (where # is the radio's address) and read three bytes (3D, one byte of mode with 3 in the leftmost nibble, and 3E).

Command 4 — offset data. This command allows the computer to read or write the DUPLEX offset. It is similar to the OW button on the 271/471. The operand is a five-digit number, the offset in kHz. The sequence:

43 4D 40 40 46 40 40 4E

i.e., 0 0 6 0 0 kHz.

would set the offset of a 271 to 600 kHz (we didn't really need the last two zeros), while the sequence:

45 4D 40 45 4E would set the offset of a 471 to 5 MHz. You can also read back an offset.

The Offset command is of dubious value. First, there's no way to specify the offset direction or turn DUPLEX on or off. DUPLEX operation can be controlled only by front panel buttons. Also, while DUPLEX values can be stored and retrieved on the 751, DUPLEX operation is not a supported feature! The DUPLEX button on the front panel of the 751 is really SPLIT — i.e., you transmit using one VFO and receive using the other.

Command 5 — set Memory/VFO. This command allows the computer to switch between a VFO and the 32 memories. It is write-only. Memory/VFO is denoted by two hex digits. The values are:

00 — VFO	11 — 17
01 — 1	12 — 18
02 — 2	13 — 19
03 — 3	14 — 20
04 — 4	15 — 21
05 — 5	16 — 22
06 — 6	17 — 23
07 — 7	18 — 24
08 — 8	19 — 25
09 — 9	1A — 26
0A — 10	1B — 27
0B — 11	1C — 28
0C — 12	1D — 29
0D — 13	1E — 30
0E — 14	1F — 31
0F — 15	20 — 32
10 — 16	

The sequence: 51 5D 51 53 5E would set a 751 to memory 19, while the sequence: 53 5D 50 50 5E would set a 271 to the current VFO. Note that this command gives you no way to switch between VFOs. You're stuck with whatever VFO you started with. Also, the command is write-only, so you can't read the current VFO/Memory — your program will have to remember it. Finally, this command is an example of the radio not really using "E" as a delimiter. If it did, 0E and 1E would not be valid operands.

Command 6 — Memory read/write. This command allows the computer to transfer information between the VFO and memory. The command is write-only and is functionally identical to the WRITE and M>VFO buttons on the front panel. The operand is a single digit:

1—VFO to Memory

2—Memory to VFO

The sequence: 63 6D 61 6E stores the information in the current memory on a 271.

a sample basic program

Figure 2 is a simple BASIC program to control a 271 and 471 for use with OSCAR-10. The program reads the 2-meter downlink receive frequency, then calculates and sets the 70-cm uplink transmit frequency. In this simple example, doppler can be accounted for by using the RIT on the 271. A more elaborate program would include automatic doppler correction computed from Keplerian orbital elements.

```
0010 REM        6522 addresses, constants, and masks
0020 DPORT = &HFF4F             ' data port address
0030 CPORT = &HFF40             ' control port address
0040 SRQ = &H04                 ' SRQ bit
0050 WP = &H02                  ' WP bit
0060 RP = &H01                  ' RP bit
0070 DAV = &H80                 ' DAV bit
0080 DDIR = &HFF43              ' data port direction register address
0090 DOUT = &HFF                ' data port lines to output mode
0100 DIN = &H00                 ' data port lines to input mode
0110 CDIR = &HFF42              ' control port direction register address

0200 POKE CPORT,SRQ             ' set SRQ high, WP and RP low
0210 POKE CDIR,SRQ+WP+RP        ' set SRQ, WP, and RP lines output, DAV input

0300 POKE CPORT,0               ' set SRQ low (begin command)
0310 Y = &H23                   ' command 2, address 3 (freq data for 271)
0320 GOSUB 2000                 ' send Y to radio
0330 GOSUB 3000                 ' read and discard the "D"
0340 FREQ = 0
0350 FOR I = 1 TO 9             ' read 9 digits of frequency
0360 GOSUB 3000
0370 FREQ = FREQ * 10 + Y
0380 NEXT I
0390 GOSUB 3000                 ' read and discard the "E"
0400 POKE CPORT,SRQ             ' set SRQ high (end command)

0450 X$ = "0" + RIGHT$(STR$(58100400-FREQ),8)    ' calculate up-link frequency

0500 POKE CPORT,0               ' set SRQ low (begin command)
0510 Y = &H25                   ' command 2, address 5 (freq data for 471)
0520 GOSUB 2000                 ' send Y to the radio
0530 Y = &H2D                   ' send delimiter "D"
0540 GOSUB 2000
0550 FOR I= 1 TO 9              ' send 9 digits of frequency
0560 Y = VAL(MID$(X$,I,1)) + &H20
0570 GOSUB 2000
0580 NEXT I
0590 Y = &H2E                   ' send delimiter "E"
0600 GOSUB 2000
0610 POKE CPORT,SRQ             ' set SRQ high (end command)
0620 STOP

2000 REM        subroutine A - write the value Y to the radio
2010 POKE DDIR,DOUT             ' set data port to output
2020 POKE DPORT,Y               ' set data
2030 POKE CPORT,WP              ' set write strobe
2040 IF (PEEK(CPORT) AND DAV) <> 0 THEN GOTO 2040
2050 POKE CPORT,0               ' clear write strobe
2060 IF (PEEK(CPORT) AND DAV) = 0 THEN GOTO 2060
2070 RETURN

3000 REM        subroutine B - read a byte from the radio into Y
3010 POKE DDIR,DIN              ' set data port to input
3020 POKE CPORT,RP              ' set read strobe
3030 IF (PEEK(CPORT) AND DAV) <> 0 THEN GOTO 3030
3040 Y = PEEK(DPORT) AND &HDF   ' get data (without command nibble)
3050 POKE CPORT,0               ' clear read strobe
3060 IF (PEEK(CPORT) AND DAV) = 0 THEN GOTO 3060
3070 RETURN
```

fig. 2. BASIC program controls ICOM 271 and 471 for use with OSCAR 10.

Since most personal computers use BASIC, the example is written in "generic BASIC." It is, however, virtually guaranteed *not* to run on your "Acceleratron-J4Q" computer without some massaging, particularly with respect to I/O port assignments. It should be fairly simple, however, to translate it verbatim to your favorite microprocessor. The example uses a memory-mapped 6522 VIA chip to exchange information with the radio. I/O addresses and constants are defined in **lines 10** through **110**; control lines are initialized in **lines 200** and **210. Lines 300** through **400** read the downlink frequency, **line 450** calculates the uplink frequency, and **lines 500** through **610** set the new uplink frequency. The subroutines at **lines 2000** through **2070** and **3000** through **3070** correspond to *Procedures A* and *B*, respectively.

extending ICOM computer control

ICOM provides a very powerful, but incomplete command set for controlling the radios. Unfortunately, the radios lack direct commands to:

- Control the DUPLEX direction or turn DUPLEX on/off.
- Switch between VFOs or read the current VFO/Memory number.
- Turn SPLIT on/off (for repeaters or HF split operations).
- Control PL frequency or turn PL on/off.
- Control filters (other than CW/RTTY-Narrow).
- Switch between HAM and GEN mode on the 751.
- Control RIT/XIT.
- Control volume, squelch, tone, RF gain, power, or noise blanker.

First, the good news: combinations of the six standard commands can be used to achieve many of the above functions. In what follows, I'll discuss several interesting functions that can be performed. Many others are possible.

Now, the bad news: what I'm about to describe is not for the meek or timid. We're talking major brain surgery — i.e., changing the contents of your radio's RAM. There are downside risks. Even *thinking* about changing the contents of this memory probably voids your radio's warranty 87 different ways. Furthermore, what I'm about to describe may not even work on your radio. There's no guarantee that the memory map — i.e., the addresses and values — for my radios is the same as the memory map for yours. Nor is there any guarantee that the memory map will stay the same in future ICOM products (or even later models of the same product). Also, a slip of the scalpel, so to speak, can leave your radio lobotimized, and in need of a brain transplant (or at least a fresh memory, available from ICOM for $25). *Proceed at your own risk!*

Before venturing further, we must delve a bit deeper into the computer architecture of the radios. From a computer standpoint, the radios look like your garden-variety, vanilla-flavored microcomputer. They have CPU, a memory, display, and an 8 x 10 keyboard. The R-71, 751, 271, and 471 — and most likely the 1271 — use the same computer architecture. In fact, they all use the same CPU chip and ROM program. The program supports all the features of all the radios. The "personality" of each radio is determined by a small 2 x 2-inch (5.08 x 5.08 cm), removable board containing a CMOS RAM. This RAM contains the radio's bands and band limits, the current VFO and memory channel, and the number of memories available, as well as the frequency, mode, band, duplex offset, duplex direction, and PL frequency for each VFO and memory. Changes to the contents of this RAM can result in drastic behavioral changes in the radio. All sorts of wonderful, unintended functions can be performed, such as switching between HAM/GEN modes and extending the frequency coverage.

The R-71, 751, 271, and 471 each have 32 memories for saving user information. The RAM to be changed is mapped into frequency and mode information for memories 33 to 255. These locations are inaccessible to the casual user operating the radio from the front panel controls. However, the memories are accessible via the computer interface.

switching bands

Each radio can cover one or more bands. The bands are expressed as upper/lower frequency bounds as in the Band Data command. The 751 has ten bands:

0	30.0 — 0.1 MHz
1	2.0 — 1.8
2	4.1 — 3.45
3	7.5 — 6.95
4	10.5 — 9.95
5	14.5 — 13.95
6	18.5 — 17.95
7	21.5 — 20.95
8	25.1 — 24.45
9	30.0 — 27.95

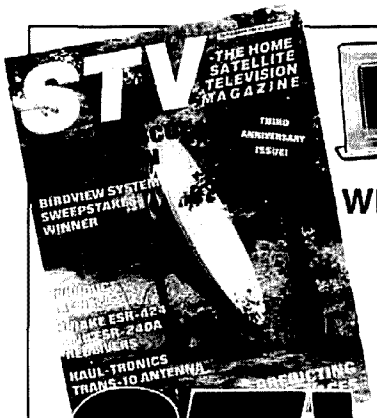
On the 751, band 0 is the General Coverage mode, while bands 1 through 9 are Ham mode. The 271 has two bands:

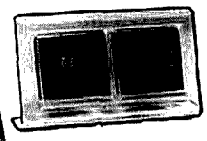
0	150.0 — 140.0 MHz
1	148.2 — 143.8

The CPU stores a single-digit band index, along with the frequency for each memory and VFO. The hundreds-of-kHz frequency digit at memory channel 38 is also the band of channel 38. To switch bands, first issue a Set Memory/VFO command with hex 26 as the operand. Next issue a Frequency Data command to write a frequency with the appropriate band index in the hundreds-of-kHz digit followed by a **M**emory **W**rite command. The frequency that you use must be within the band limits of the radio. Also, be very careful in selecting the band index, as an invalid band will result in an upper/lower frequency pair of 0.00 through 0.00 Hz, not a very useful pair! Finally, issue a Set Memory/VFO command with 00 as the operand to return the radio to the current VFO followed by a **M**emory Read command. The bands of Memory channels 1 through 32 can also be changed through memory manipulation, but, in general, it's easier to change the band of a VFO and then store the VFO in a memory than it is to change the band of a memory directly.

greater frequency coverage

The frequency coverage of several of the radios can be extended beyond the band limits. For example, many 751s can be tuned below 100 kHz and above





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30.0 MHz. The additional frequency coverage appears to be both model- and radio- dependent. It is sometimes even mode-dependent.

All frequencies entered via the dial, the keypad, or the computer interface are checked to ensure that they are within the band limits. Frequencies outside the band limits are rejected. However, there are several ways to evade this check. The simplest is to find a frequency in memories 33 through 255 and transfer it to the current VFO. The radio does no checking on memory/VFO transfers. Frequencies both above and below the band limits can be obtained. Frequencies above the band limits can also be generated by using the hex digits "A," "B," "C," and "F" in certain digit positions in the Frequency Data command. Use of these digits generates a carry in the next higher digit.

In general, whenever you are outside the radio's band limits, you may tune only *towards* the band limits. For example, if you are above the band limits, you may tune only lower in frequency. If below, you may tune only higher in frequency. Attempts to tune in the opposite direction will place you at the opposite band limit.

conclusion

ICOM has incorporated computer control into its radios, and its current product line continues that innovative trend. With minimal hardware, any microcomputer can be used to control the radios. The standard command set is simple, easy to use, and sufficient for most applications. Many functions not provided for by the standard command set can be realized by combinations of commands. Although it's not discussed here, it's also a simple task to intercept the radio's keyboard matrix and simulate button pushes with a computer. There's no question that computer-controlled radios can take the drudgery out of, and put the fun back into, Amateur Radio operation.

ham radio

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Prevent spikes from destroying your equipment

AC line transient protection

It all started when I destroyed my VCR with a lawnmower. Perhaps I should explain in greater detail. One summer day as I was trimming the weeds in my *front lawn, my electric lawnmower blew a motor* field rectifier and began to draw considerable current from the AC line. This situation was quickly, but not instantly, corrected by the house circuit breaker, which tripped and broke the circuit, as it's intended to do. Then, however, the electric fields in the motor collapsed, producing a large back EMF on the now-open circuit AC cord. Unfortunately, this cord connected the mower to the same circuit on which my VCR was patiently awaiting the start of a "Sky King" rerun.

The VCR, like a lot of new equipment — including most new Amateur rigs and the computer on which I write — doesn't like to see high voltage spikes coupled into its relatively fragile CMOS logic integrated circuitry. On a computer, line spikes or transients can cause data drop-outs, so-called "soft" errors in the RAM memory, or other grief. In my VCR, the unusually large transient simply fried some component on the microprocessor board. Fortunately, the VCR was still under warranty.

I decided that some transient protection would be necessary. A quick check of the catalogs showed that many manufacturers make line cord sets with transient suppression to reduce the chances of just this sort of occurrence. But these outlet boxes cost at least $30.00, so I decided to try to build some lower-cost version of these outlet boxes, using the same type of line transient protection devices. I could then distribute these protective boxes about my house to protect any electronic equipment that would be sensitive to high voltage transient peaks on the AC mains.

In this project, I've taken a low-cost approach to transient protection that uses commonly available metal oxide varistors. The method could easily be adapted to many of the commercially-available outlet boxes or AC junction strips found in most hardware stores. The total cost of each protected outlet box is about $5.00.

metal oxide varistors

A metal oxide varistor (MOV) is a voltage-dependent semiconductor device that acts much like a pair of back-to- back zener diodes. The MOV is placed across the AC input of the device to be protected. Under normal conditions, the MOV has a high input impedance so it draws a minimal amount of power. However, if the voltage across the AC line increases to a point above the turn-on voltage of the MOV, it suddenly switches to a low-impedance state. This low impedance is in shunt with the line, so the AC voltage is limited. Once the transient passes — i.e., the *AC line voltage returns to normal* — the MOV recovers and returns to its high-impedance, stand-by state.

Most of the voltage increases that appear on the AC mains are of momentary duration and are caused by switching large loads, especially inductive loads, on or off, or by lightning strikes (at a distance, not directly on the equipment). This means that even though the transient may be many hundreds or even thousands of volts, since it does not last long (a "typical" transient might last a few tens of microseconds) there is little total energy in the spike, and the energy can be safely dissipated in the MOV.

A typical MOV intended for use on a 120 volt AC line has the specifications shown in **table 1**. It can take up to 4500 Amperes in a spike with a total energy of 35 Joules, which means that if it is clamping the line at the specified maximum of 225 volts, the transient may last only about 30 microseconds. This explains how a small device, no bigger than a rather large disk capacitor, can tolerate 4500 amperes — it does so only for a few microseconds, and not too often, at that. However, by clamping the line to 225 volts, the MOV can do a great service to us in protecting the seemingly delicate, certainly complex equipment we now find commonplace in our homes, computer rooms and ham shacks.

By Jerry Hinshaw, N6JH, 4558 Margery Drive, Fremont, California 94538

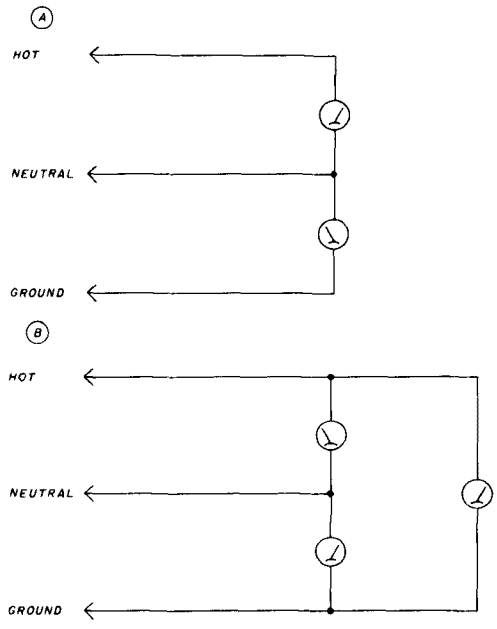


fig. 1. An illustration of the two modes of voltage transient which are encountered on a typical 3 wire circuit. (*A*), the case when the potential between hot and *neutral* is abnormally high due to a transient. (*B*), an alternate transient mode where the power-carrying lines are elevated above the safety ground.

The manufacturers' data sheets provide detailed design information that permits us to calculate the type of MOV needed for any expected transient. However, I found such data to be of little practical use because I can't predict what type of transient one might expect on the AC lines in my typical tract home. Therefore, I merely selected a MOV that had a large current capacity and the lowest available varistor voltage because these factors seemed likely to offer the maximum protection. Furthermore, there was only one type available at the local Radio Shack, which pretty well directed my choice. In general, though, it's best to select a MOV with the lowest turn-on voltage that will still permit normal operation of the equipment you choose to protect. This choice means that a maximum range of clamping is available and that the minimum excess spike is coupled into the equipment. More detailed information is available from the manufacturers.

There are two modes of voltage transient which the suppressor should be able to shunt. In the first, a transient may cause the voltage across the AC lines to rise above the nominal 120 volts rms (I am speaking about the US standard mains, but these concepts apply to power mains worldwide). In this case, the voltage between the two lines rises abruptly above the nominal value with respect to a ground reference. This can be called a differential mode transient because there's a difference between the reference point and each of the two lines.

Figure 1A shows this common mode voltage transient schematically, with two hypothetical voltmeters placed across the AC lines. These meters are hypothetical in that they are presumed to have instantaneous response and are depicted at the exact moment a transient has caused the voltage across the hot and neutral lines to soar far beyond the nominal value. However, the neutral-to-ground voltage is not significantly disturbed.

The second type of transient is shown in **fig. 1B**. Note that the potential across the hot and neutral lines can be normal, or nearly so, while their potential to ground can be very great. This type of "elevation" can be as damaging to electronic circuitry as the differential type of transient. In order to protect equipment from these two distinct modes of transient behavior, we need to have two sets of suppressors. One suppressor must be placed across the AC line and a second set should be placed from each side of the line to ground **fig. 2B**. Clearly, if there is no ground line present, as in **fig. 2A**, only one MOV suppressor is needed for protection, as long as no other path to a safety gound exists.

table 1. Specifications of a typical MOV designed for 120 VAC use, the General Electric V130LA10A.

Varistor voltage, minimum	185 volts
Varistor voltage, nominal	200 volts
Varistor voltage, maximum	225 volts
Peak current, maximum	4500 amperes
Energy	35 joules

construction

Now that the MOV has been selected, it must be safely installed across the AC input of the equipment to be protected. One good way to do this is to install it inside the equipment itself. Another way is to somehow place the MOV across the input line between the equipment and the AC outlet. I chose to do the latter because I found a source of low-cost AC outlet boxes that suited my needs.

The AC box I chose is a plastic unit originally designed to expand a US standard two-plug outlet into a six-outlet box. It is easily installed by removing the usual switchplate and plugging the new unit into the wall. A single screw secures the outlet box into the threaded insert that originally held the outlet cover

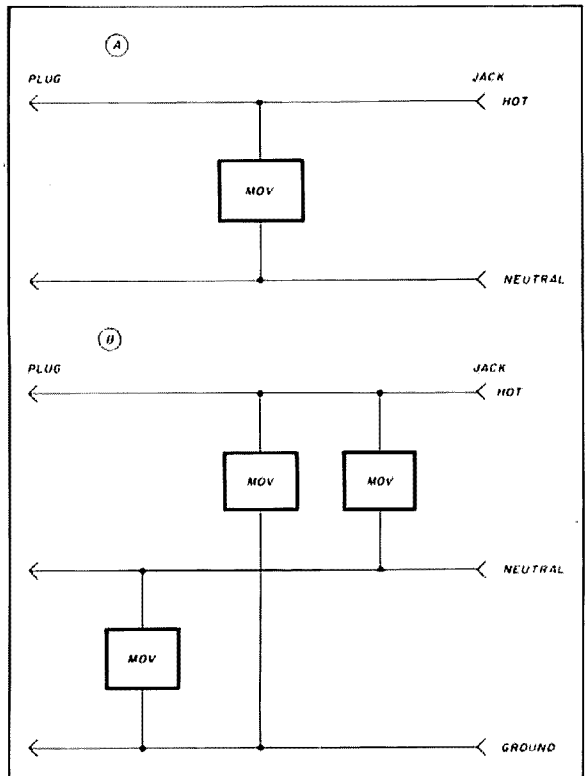


fig. 2. A simple schematic diagram showing how the MOV is installed to protect equipment from power line transients. (*A*), placement of the MOV in a two-wire system. (*B*), Three MOVs are used to fully protect a system which has a third conductor as safety ground.

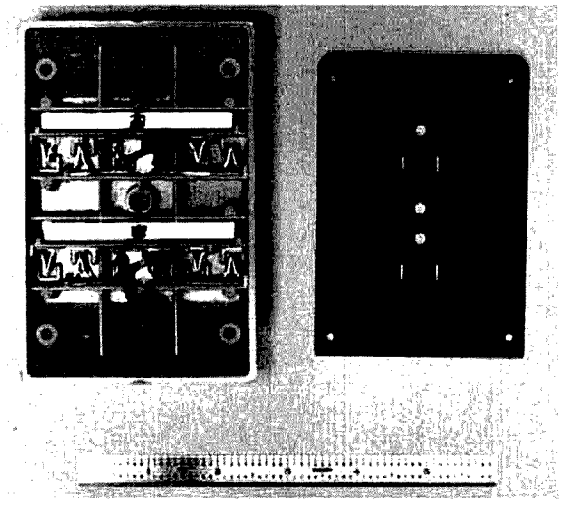


fig. 3. Interior view of the accessory outlet box with one MOV installed across the AC hot and neutral lines. Note the protective sleeving on one of the MOV's leads.

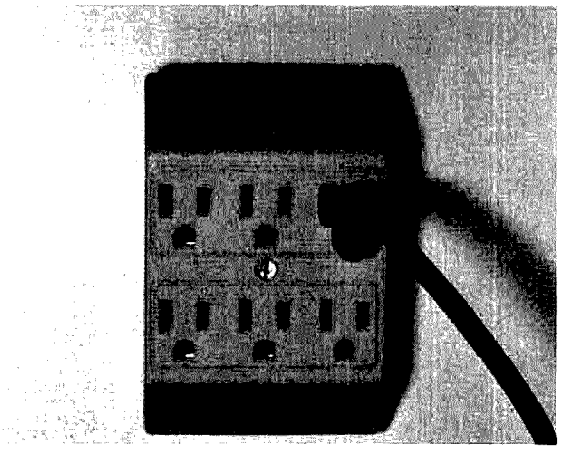


fig. 4. The completed outlet box installed and in use.

plate. This type of accessory box seems widely available and has the advantage of providing extra outlets, which are often useful. Furthermore, it has enough room inside to mount several MOVs easily. You can separate the box into two independently protected circuits, if you desire.

The outlet box I chose was designed to convert two standard three-wire outlets into six three-wire outlets. Made of plastic, it contains two sets of contacts, each set consisting of three conductors, one each for the ground, neutral and hot sides of the AC line. The interior construction of the box is shown in **fig. 3.** The box protrudes perhaps 1.5 inches out from the wall when it's installed, so there's space inside for installing the MOVs.

The figure also shows one MOV installed across the two main conductors of one set of three outlets. No protection against common mode transients has been installed, so there's no MOV installed between the ground pin and either of the other two pins. This is because in my house, although some of the outlets do have ground pins, there are, in fact, no ground connections actually present inside the wall. The original outlets in this house are two-wire outlets, which is typical of most houses built before the 1960s, when local electrical codes gradually began to require the use of three-wire, grounded outlets. Thus, there's only one MOV installed in this box. This MOV is placed across the hot and neutral lines as shown. The metal conductors easily take regular tin-lead soldering, so that it is simple to permanently solder the leads of the MOV across the circuit. Note that one of the leads of the MOV must cross over the other lead's AC connection. At this point it is imperative that you install good insulation. I used two pieces of heat-shrink tubing, one inside the other. Remember, this circuit is not powered by a low-voltage supply like a typical digital breadboard — think "safety" throughout the project!

Once the MOV circuitry is installed, replace the back cover of the box. This particular unit is installed after the cover plate of the old outlet has been removed. A long screw secures the box to the wall. The completed unit, ready to install on a wall outlet, is shown in **fig. 4**. It protrudes a bit from the wall, but where there were two unprotected outlets before, there are now six transient-protected outlets available. The total construction time, if you want to call such a simple procedure by so elaborate a term, is well under an hour.

Now my VCR, my ham shack, and my computer each have their own transient protection box. The lawnmower has yet to blow another motor rectifier, but I have fair confidence that, should it happen again, the delicate electronics components will be better off than before, when they faced high voltage transients completely unprotected. The cost of a few of these suppressor outlet boxes seems like cheap insurance to me.

ham radio

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SPECIFICATIONS

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Memory	32K EPROM, 16K RAM standard
HDLC	Packets are controlled by hardware for maximum performance permitting full duplex operation
Modem	1200 baud Bell 202 compatible (standards easily configured for 300 baud 200 Hz shift for HF use
Serial	Computer/Terminal port is industry standard RS 232 C compatible for use with most equipment
Radio	Watchdog timer for channel protection from transmitter audio levels adjustable for nearly any radio. Wide dynamic range demodulator. Channel busy input (IR, DCD) to inhibit packet transmissions on a shared channel
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- eight additional memories
- 10-Hz readout
- scanning
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modifying the Trio-Kenwood TS-930S

Kenwood's TS-930S transceiver includes a number of unadvertised capabilities. This article describes four of them that can be enabled by making just four simple modifications. These modifications require no additional parts except for one solder lug and about 3 feet (0.9 meter) of No. 18 (or smaller) insulated wire. They can be completed within an hour after removing the 930's covers.

I'll describe the modifications first, then explain how to install them.

four simple mods

- **Mod 1: eight additional memories.** Adding one jumper results in each VFO (A and B) having 8 memories, creating a total of 16.
- **Mod 2: 10 Hz readout.** Ever notice the unused seven-segment LED on the left end of the frequency display? One jumper makes it usable by shifting the frequency display one digit to the left, resulting in 10 Hz resolution of the displayed frequency.
- **Mod 3: Scanning.** Add one jumper and the 930 will scan through the 8 frequencies stored in either of the VFO A or B memories.
- **Mod 4: Full coverage on transmit.** Add three jumpers and the 930 is ready to transmit on WARC, MARS, and the remainder of the *nonamateur* frequencies in the 1.5–30 MHz range.

getting ready

As with all modification articles, please read this article several times before you heat up the soldering iron. Doing so may well save you headaches later when you apply power to the set.

After disconnecting everything from your 930, remove the top and bottom covers (16 screws) and place the rig top side up, facing you, on a cushioned surface. Each of the modifications requires access to the digital-unit board, which is hidden under the speaker and VOX control assembly, which can be removed by removing the four screws that hold it to the main chassis of the 930. Lift the assembly upward slightly and disconnect the small 2-conductor plug (with the red and white wires) from the digital-unit board. Disconnect the speaker leads (remember their polarity) and the other two connectors that plug into the small board directly beneath the VOX controls. Set the assembly aside.

Two of the mods require access to the back of the front panel. This is easily accomplished thanks to the cabinet's sensible design. On each side of the 930 you'll find the front panel mounting brackets. There are two flathead screws and one roundhead screw in each bracket. Refer to **fig. 1** for their locations. Move the 930 toward the front of your work table so that a few inches of the rig hangs over the edge. The panel will tilt forward after (1) removal of the two flathead screws from each bracket and (2) careful loosening — *not removal* — of the roundhead screws. The panel may tilt on its own, so keep one hand on it while you

By Roger J. Hoffman, WB9BXT, 5719 La Vista Drive, Alexandria, Virginia 22310

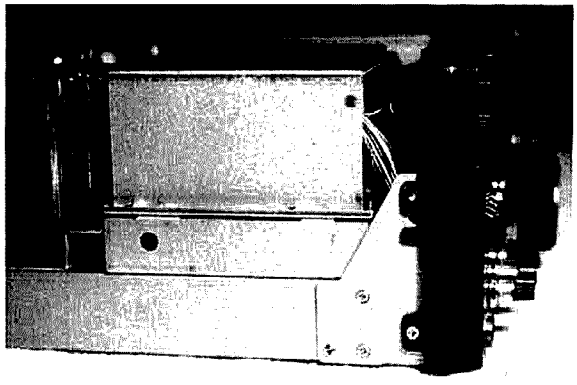


fig. 1. Front panel mounting bracket screws.

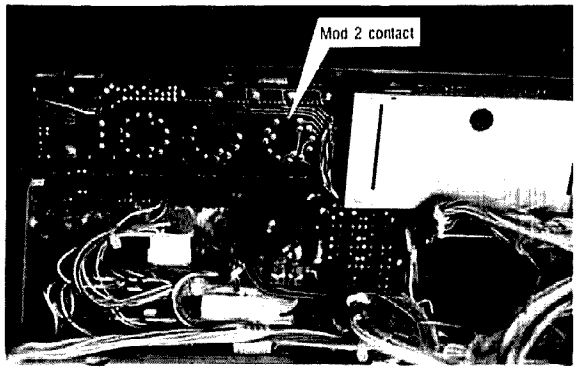


fig 2. Contact for mod 2 is at 12 o'clock position.

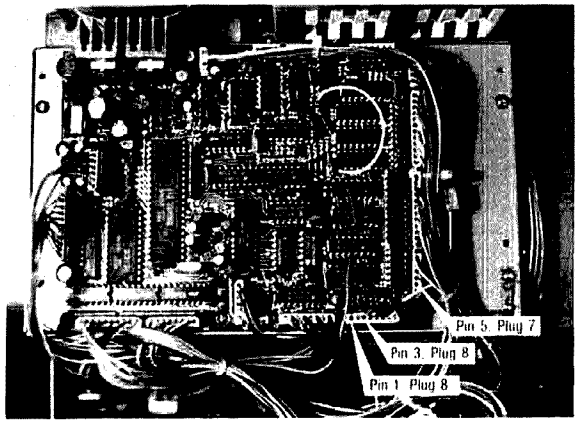


fig. 3. Connections for mods 1, 2, and 3 on digital unit board.

loosen the roundhead screws. Tilt the panel down about 60 degrees and retighten the roundheads. This will help to maintain the tilt of the panel and will prevent straining the multitude of wires connected to it.

mod installations

Mod 1 (8 additional memories) requires a jumper from pin 5, plug 7 on the digital-unit board to ground through a switch. The function switch (VFO A, VFO B, etc.) has an empty contact to ground when it is placed in the VFO B position. **Figure 2** shows the location of the switch contact on the back of the switch's circuit board. Check continuity to ground through this contact to verify that you have the right one. Remember to place the switch in the VFO B position for this check. Solder one end of the jumper to this contact. The other end of the jumper needs to be bent into a small hairpin loop and fitted into the empty hole for pin 5, plug 7. (See **fig. 3** for the location of plug 7.) My 930 is seldom moved, so I don't worry about the jumper possibly pulling out of the hole. You'll need to experiment a bit with the size and shape of the hairpin to achieve a snug fit.

When finished, power up the 930 and program a few frequencies into the VFO A memories as you normally would. Then select VFO B and program a few more frequencies. Recall the memories, switching between VFO A and B. You'll notice that you now have the capability of 16 memories. If not, go back and check your jumper.

Mod 2 (10-Hz readout) requires installation of a jumper from pin 1, plug 8 on the digital-unit board to ground. (See **fig. 3** for the location of plug 8.) I used the hairpin trick again to connect the plug end of the jumper. A solder lug is connected to the other end, which can then be connected to any convenient screw in the chassis. I used one of the speaker/VOX assembly holddown screws. Test the mod by powering up the rig.

Mod 3 (scanning) requires a jumper from pin 3, plug 8 on the digital-unit board to ground through a switch. Use the hairpin method to connect the jumper to the plug. The other end of the jumper connects to the panel light DIM switch, which has an extra contact to ground when it's in the DIM position. (**Figure 4A** shows the location of this contact, with **Figure 4B** showing this in greater detail. Again, check continuity to ground with the switch in the DIM position to verify that you have the correct contact. When you've finished this modification, power up again, load up the memories, select VFO A and depress the DIM switch. Notice that the scanning starts with memory channel 1, scans to 8, and repeats, stopping on each channel for about 2 seconds. To scan VFO B memory channels, you must first initiate scanning in VFO A and then select VFO B. Scanning will not initiate in VFO B. In addition, only 8 channels can be scanned (that is, either VFO A or VFO B).

Mod 4 (full coverage transmit) requires three jumpers on the digital-unit board. The first one provides transmit coverage for the WARC bands. The other two provide the remaining coverage. If the WARC jumper is not installed, the 930 will still trans-

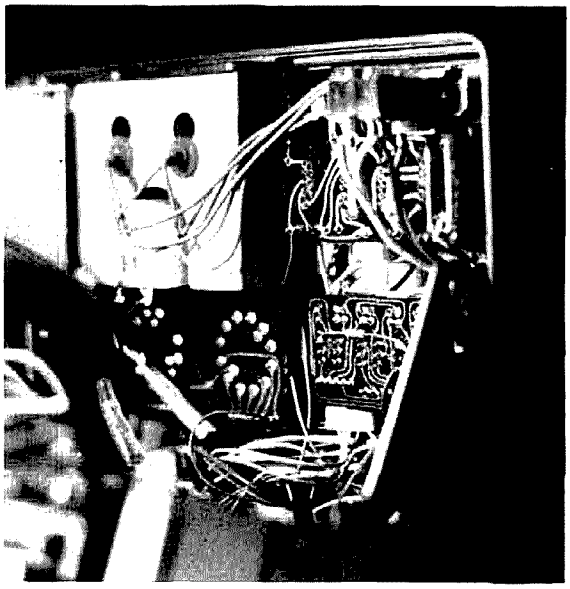


fig. 4A. Dim switch circuit board.

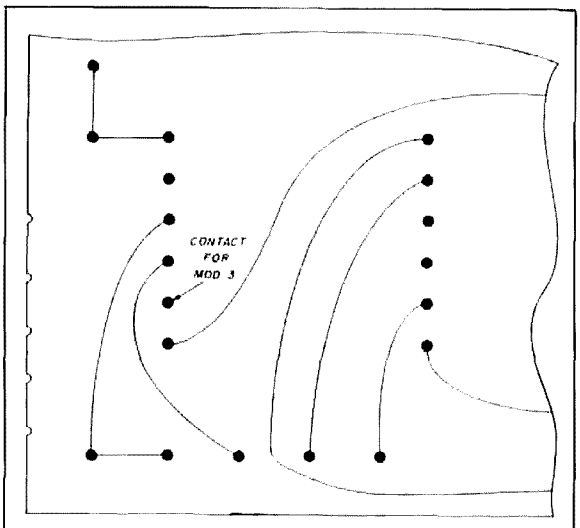


fig. 4B. Lower left corner of dim switch circuit board.

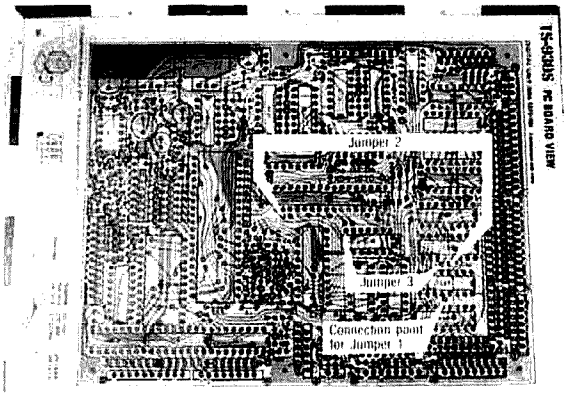


fig. 5. Location of connection points for mod 4.

mit over the entire 1.5 to 30-MHz range with the exception of the 0.5-MHz segments, that contain the WARC bands. If you want only the WARC coverage, install only the first jumper, which goes from pin 12, U23 to ground. (Note: on two of the three 930s I've modified, the WARC jumper had already been installed at the factory). The second jumper goes from pin 9, U11 to pin 12, U21. The third jumper goes from pin 9, U12 to pin 12, U22. A close inspection of **fig. 5** will show that each of the connections to the above ICs can be made on unused solder pads on the digital-unit board. I melted a small amount of solder on each of the pads before installing the jumpers. The grounded end of the WARC jumper can be attached to the same solder lug that was used for the 10-Hz mod. You'll find that the optional tuner (AT-930) covers the WARC bands, but not the general-coverage bands.

final steps

Reinstall the speaker/VOX assembly, remembering to reconnect the four cable assemblies that were disconnected earlier. Reattach the front panel, taking care not to pinch any wires. Replace the covers, and enjoy!

conclusion

What's my assessment of the mods? Well, I hardly ever used the eight memories that came with the 930, so I really didn't need eight more, although I do use some of them now for scanning. I use the scanning feature to locate the family net at 14.177 MHz (±) by programming from 14.175.5 to 14.179.0 in 50-Hz steps and scanning through them while attending to other tasks in the station. I also use it for checking band openings by programming frequencies in different bands. One caution: the 930 will scan as long as the DIM switch is depressed. This includes the transmit mode, so be sure to disable the scanning before transmitting! The 10-Hz resolution isn't needed except to program scanning frequencies, so it's really just a novelty. The full transmit coverage is necessary if you want to use the 930 on some of the MARS frequencies, as I do.

Thanks go to DL3AM and KW9G (ex-WA9GMK), who assisted in installing these modifications. Thanks also to Trio-Kenwood for its courteous approval of my request to reproduce portions of the 930 Technical Service manual for this article. Copies are available from TRIO-KENWOOD, 1111 West Walnut Street, Compton, California 90220.

ham radio

PRACTICALLY SPEAKING ... Joe Carr K4IPV

keep it cool . . . and you'll keep it long

"Keep cool" — good advice for people on a hot day, and good advice for electronic equipment anytime. Heat is the number-one assassin of electronic equipment.

Many device ratings are based on maintaining certain operating temperatures. One manufacturer of a "hobbyist grade" audio power transistor, for example, offers (and advertises prominently) a transistor with a seemingly tremendous collector power dissipation. But there's a catch: the power is available only at room temperature (77-86 degrees F, or 25-30 degrees C). At temperatures above 30 degrees Celsius, the transistor must be derated substantially. No matter where the transistor is used, if it's inside a cabinet or box the temperature will almost certainly exceed 30 degrees C!

Similarly, RF power transistors in transmitters die as often from overheating as from that elusive gremlin, VSWR, but the problem is less well recognized. I know one ham who lost the power transistors in his trunk-mounted 100-watt 2-meter power amplifier several times before he realized that the heat was the culprit! During the summer months, the trunk of a car will sizzle even though the air-conditioned passenger cabin cools off within a few minutes. Moving the amplifier to behind the dashboard cured the problem.

Reliability experts measure equipment performance in terms of "Mean Time Between Failure" (MTBF), which is usually expressed in hours. For example, an MTBF of 1000 hours implies that, for a large number of samples of the equipment, an average of one soul-destroying failure per thousand hours of operation will occur. One source claims that a 10-degree C rise in operating temperature will cut the MTBF almost in half.

Just how important is cooling in electronic equipment? Let's consider some examples. About ten years ago

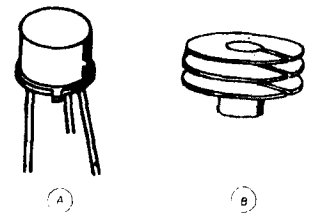


fig. 1. (*A*) Typical low-signal (low-heat transistor) does not use a heatsink. (*B*) Top-hat finned heatsink helps power transistors run cooler.

I worked in a university hospital, repairing patient-monitoring equipment. The EKG oscilloscopes at the nurses' central station were a reliability nightmare. About once a week, usually at 3 AM, the staff would call me to come repair one of the four 'scopes. Yet the same model 'scopes operated reliably at the patients' bedsides. The problem was overheating of the central station 'scopes, which were mounted inside a completely closed desk/console. After ten 1-inch ventilation holes were cut and a pair of 100-CFM "whisper fans" were installed, central station 'scopes became as reliable as the bedside 'scopes.

A second example is a story of tragedy prevented. My first personal computer was a Digital Group, Inc., Z80-based machine with 26K (2102 chips) of static memory. In those days, my kilobuck bought (in kit form) a motherboard, three 8K memory boards, a CPU board, a 64-line TV/cassette interface board (with some static memory chips on-board), and several input/output boards. All of those boards contained lots of TTL devices, and they generated a large amount of heat. The builder had to supply the cabinet, a ±12 VDC, 1-ampere dual-polarity power supply, and a +5 volt DC, 10-ampere regulated power supply. Since I operated the computer next to a ham rig, EMI both to and from the computer was an issue, so I had to use a well-shielded aluminum cabinet—and shielding isn't always compatible with heat dissipation.

At first all those cards and the two DC power supplies were buttoned up inside the almost unvented aluminum cabinet. Needless to say, the temperature of the cabinet rose to egg-frying levels, and the HEP S-7000 power transistor used as the series-pass element in the voltage regulator operated hot enough to take off skin when touched. I knew that computer would be a reliability headache if the heat were not removed, so I installed a pair of 40-50 CFM fans: a 3.5-inch (8.89 cm) model blowing across the S-7000 heatsink and a 4.5 inch (11.43 cm) model cooling the printed circuit board compartment. Because of the EMI

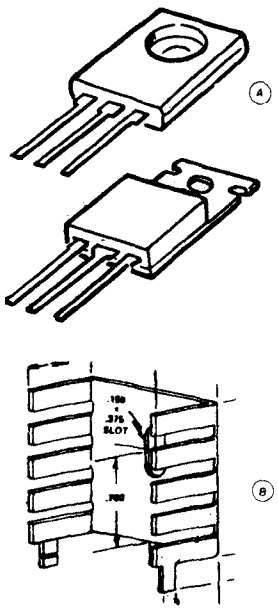


fig. 2. (*A*) Two forms of plastic power devices packages. (*B*) Vertical or horizontal finned sheet metal heatsinks for the above devices.

problem, the ventilation and blower opening were covered with perforated aluminum sheet metal.

No one with any electronics experience — however slight — can deny that heat is the primary killer of electronic devices. Projects or equipment that pass or deliver large amounts of either current or power must be kept cool for proper operation. The methods given in this article are simple and should be sufficient for most reader's applications. While reliability engineers and thermodynamicists may flinch at the lack of mathematical elegance, the methods are nonetheless effective.

There's only one simple rule: where there's excessive heat, remove it. What do I mean by "excessive?" If the equipment feels too hot to touch, or has a history of unexplained failures and/or repairs, then it's probably running too hot. An engineer will have specifications to meet and calculations to make, but these are beyond the scope of this article. The empirical "skin of the thumb" rule, however, suffices for our needs.

Three basic tactics can be used either singly or in combination to dissipate heat:

- radiate more heat,
- improve natural ventilation, or
- add or increase forced-air cooling.

For most readers, water cooling isn't relevant even though some commercial broadcast transmitters use circulating water for cooling. In fact, I once worked in a 10 kW AM broadcast station that used the waste heat from the transmitter's water radiator to heat the transmitter building!

protecting transistors and IC regulators

On small projects where it's not practical (or possible) to use forced-air cooling, you'll have to provide heatsinking for the semiconductors. In fact, even most forced-air cooled projects will need these metal radiators. **Figure 1A** shows the metal TO-5 transistor package. Most of these transistors are mounted on printed circuit boards and are low-signal (and low-heat) devices. But certain TO-5 transistors operate at moderate power levels (in audio drivers, for example). A "top-hat" finned heatsink (**fig. 1B**) is mounted on the TO-5 package to radiate heat. There are also other "spring clip" versions of this same kind of heatsink.

Figure 2A shows two forms of plastic power device package. You'll find these packages in power transistors (e.g., 2N5249), thyristors, and three-terminal IC voltage regulators. In the

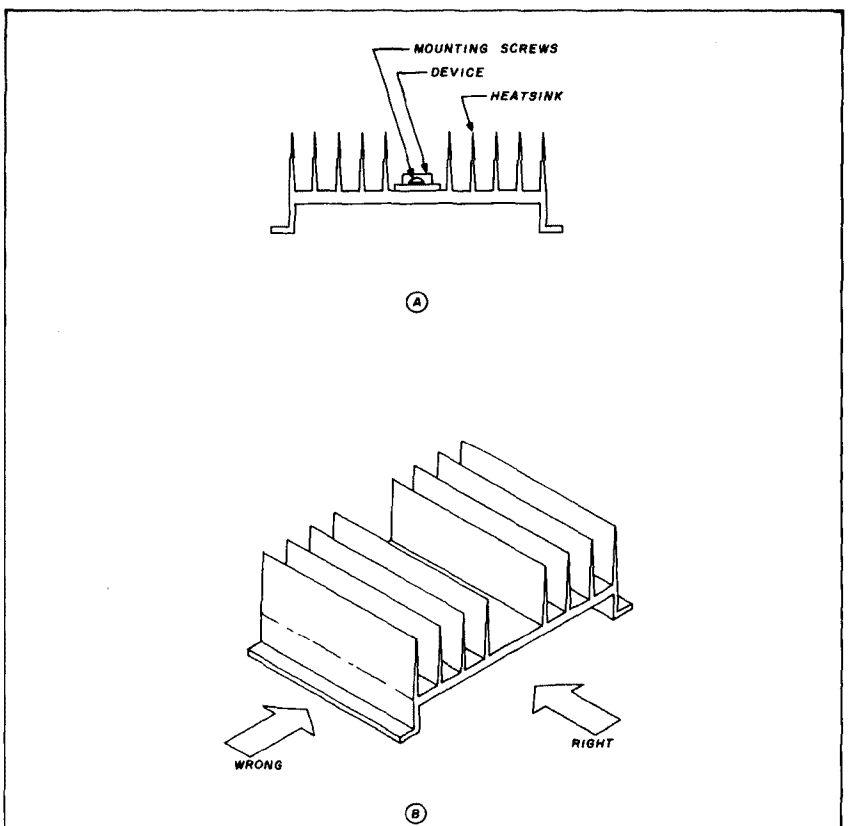


fig. 3. (*A*) Large finned heatsink used with TO-3 transistors, high-current voltage regulators, high current diodes and SCRs. (*B*) The "right" and "wrong" way of forcing air over finned surfaces.

case of regulators, the devices are often rated at 750mA in free air and 1000mA when heatsinked. Either vertical or horizontal finned sheet metal heatsinks (**fig 2B**) are used to provide heat dissipation. Be sure to use a thin layer of silicone heat transfer grease between the metal tab surface on the transistor (or regulator) and the heatsink. Also be sure to tighten the mounting screw properly in order to facilitate heat transfer to the heatsink.

Sheetmetal heatsinks are used for TO-3 transistors and three-terminal regulators that are mounted on printed circuit boards. The bent sheetmetal heatsinks are good for up to about 10 watts of power, or voltage regulators up to 1.5 amperes. For the 3-ampere, 5-ampere, and 10-ampere voltage regulators that also use a TO-3 package, it would be better to use a larger finned heatsink.

Often the metal chassis itself is used for heatsinking. In these cases the transistors are bolted either directly to the metal chassis or mounted with mica insulators. In both cases, silicone heat transfer grease is used between the semiconductor device and the chassis. This method is especially successful when the chassis is large or unusually thick.

Some printed circuit boards use large areas of unetched copper foil and/or large metal ridges or blocks to provide better heatsinking. This method is used especially where there are not single particular devices that can be individually heatsinked (e.g., a TO-220 transistor), but rather when there are a large number of heat-producing devices (such as TTL ICs).

There are many different forms of large, finned heatsinks used for TO-3 (and other) transistors, high current voltage regulators, high current diodes, and SCRs; **fig. 3A** shows a side view of one of these heatsinks. In this case, the TO-3 transistor (or other device) is mounted on the flat central surface of the heatsink with screws. In most situations, it's wise to use a thin smear of silicone heat transfer grease between the device and the heatsink. This grease is especially necessary when a mica insulator is placed between the semiconductor device and the heatsink. Again, it's essential to make sure that the mounting screws are cinched down tight enough to allow maximum heat transfer (but not enough to distort the device package). The big concern in selecting a heatsink is the amount of surface area, measured in square inches or square centimeters.

When forced air is used to cool a heatsink — always a good idea when the power and/or current is high — then the orientation of the heatsink with respect to the airflow is sometimes important. **Figure 3B** shows right and wrong ways to force air over the finned surfaces. Keep in mind, however, that the orientation is not always critical, especially when air from the "wrong" direction is suffi-

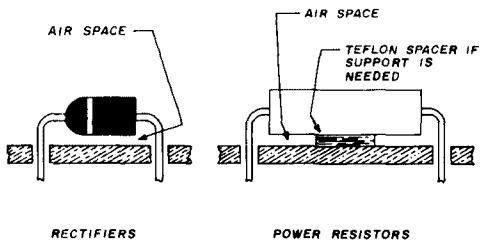


fig. 4. Suspending components above boards aids in air circulation and subsequent cooling.

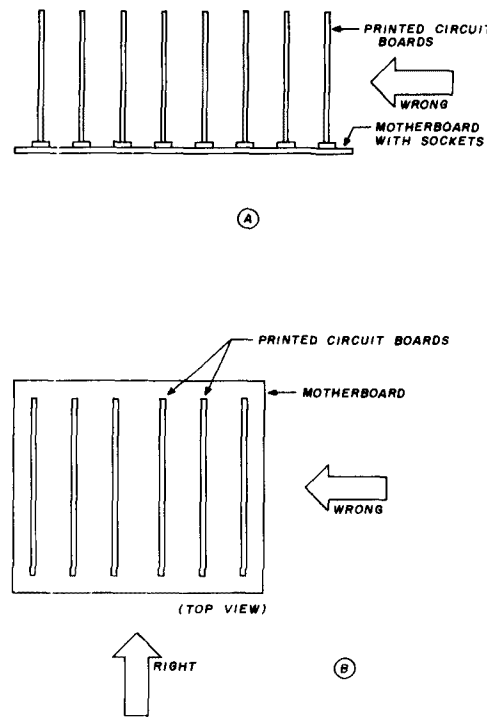


fig. 5. Correct way of forcing air past multi-bands. (A) side view (B) top view.

cient or blows over the entire surface. The designations "right" or "wrong" are merely general considerations for some critical applications.

other components

Not only power transistors generate heat. Rectifier diodes and power resistors should be mounted with their bodies 0.125 to 0.250 inches (0.317 to 0.635 cm) off the printed circuit board (see **fig. 4**). This procedure allows the heat to dissipate into the air instead of into the PCB material. I've seen many phenolic and some fiberglass printed wiring boards badly damaged from the effects of a 10-watt power resistor mounted flush to the surface. Some "bargain basement" rectifier diodes can meet their rated forward current only when the rectifier is mounted 0.50-inch (1.27 cm) off the board and has its axial leads cut to 0.75-inches (1.9 cm) or longer. Those diodes are overrated and should be used only in lower current applications or shunned entirely.

Layout is important when power components are mounted on the PCB. Try to avoid clustering power components in one small area of the board, especially when using cheap phenolic board material. Avoid placing heat-sensitive parts near power components. For example, 10-watt resistors should not be mounted adjacent to polystyrene capacitors or small transistors.

Besides reducing the operating life or limiting the power output of circuits, overheating can also decrease performance in other ways. Certain circuits — oscillators, for example — are inherently sensitive to heat. There was once a popular three-band kit-form HF transceiver that suffered immense VFO drift because the JFET VFO was located right next to the RF/IF strip tubes. Although that was such a bad design error that nothing would really "fix" the situation, a lot of Amateurs were able to improve the frequency stability markedly with some thermal insulating material placed between the RF/IF PCB and the aluminum VFO housing.

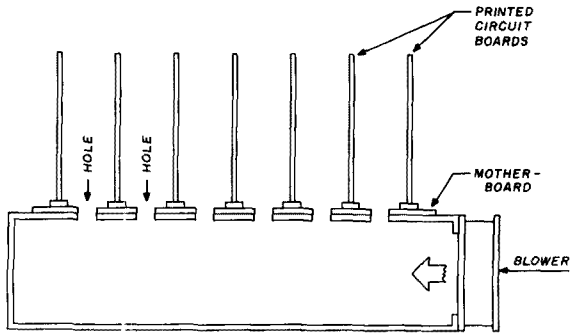


fig. 6. Another method of cooling multi-boards requires slots to be cut in socket-mounting chassis.

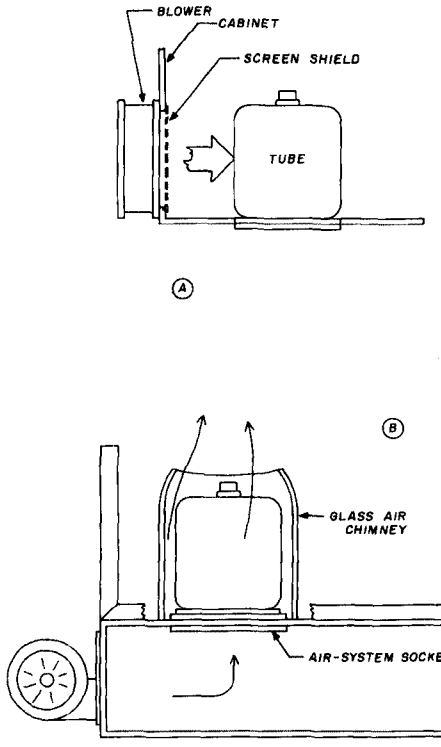
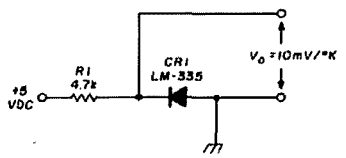


fig. 7. Several methods available for cooling RF power amplifier tubes. (*A*) Airflow is directed across glass envelope by fan. (*B*) Air flow is forced through socket and glass air chimney.

large multi-board projects

When I first felt the temperature of my Digital Group, Inc., cabinet I took steps to get rid of the heat, and reliability was improved. Rarely does the homebrew builder have the flexibility that I had with my Vector Electronics S-100 cabinet. In most cases, the builder must make do with only a single fan and must be clever to make best use of it. **Figure 5A** shows a typical large-scale multi-board project —



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

fig. 8. Temperature sensing circuit provides voltage proportional to circuit temperature.

such as a microcomputer or transceiver — in which plug-in printed wiring boards are installed on a socketed motherboard. Usually, these PCBs are mounted in a closed cabinet for reasons of both EMI shielding and aesthetics.

If we apply air broadside to the PCBs, then only the first board in the line-up will benefit. **Figure 5B** shows a top view that allows visualizing right and wrong airflow directions. Obviously, air coming in from the sides is able to remove heat from more of the PCBs with greater efficiency.

Figure 6 shows a method used by a friend of mine who built a home-brewed 6502-based computer. He used a large metal chassis with a motherboard mounted on it to hold the PCBs. He cut 3/4 inch (1.9 cm) holes in both the chassis top and the motherboard to admit air between the boards. (Although only one hole is shown between each board in this side view, there were four per row in the actual project.) Air from the blower flowed up through the holes and across the electronics components on the PCBs.

Linear amplifiers and high-power transmitters pose special heat problems. Some linears, for example, are only 45 percent efficient. Therefore, a 1000-watt linear amplifier delivers 450 watts of usable RF power and 550 watts of waste heat. To make matters even worse, the necessity of keeping harmonics at home means we button up all that heat in a shielded metal cabinet.

Most RF power amplifier tubes used in Amateur Radio equipment must be cooled with forced air in order to realize their full ratings (some are absolutely dependent on cooling). **Figure 7** shows two methods for providing the needed cooling air. In **fig. 7A** we see a situation in which a blower is mounted so that the air flow is directly over the glass envelope. The fan may be mounted either outside the RF compartment (as shown) or inside, as in the Heath SB-221.

The other method, shown in **fig. 7B**, assumes the use of "air system" tube sockets. A blower or fan supplies air to the bottom side of the socket and the air is directed upwards through holes in the socket and around the glass evelope. A "chimney" aids in keeping the airflow against the glass. Some air system sockets have plumbing connections for the air hose, while others depend upon pressurization of the lower compartment. In either case, this socket is better because the pin seals with the glass are kept cooler.

The plate cap pin seal should also be kept cool, if possible. Toward this end, some builders use a finned "heat dissipating" plate cap to make electrical connection to the anode.

temperature measurement

In some cases we'll want to provide either continuous for temporary monitoring of the actual operating temperatures. Although there are elegant methods using thermocouple junctions, we can use a simple, low-cost PN junction temperature sensor. National Semiconductor and others manufacture such devices. **Figure 8** shows the simplest circuit for the National Semiconductor LM-335 diode device. The LM-335 will measure temperature over the range −10 to +100 degrees F (−23 to +38 degrees C). In the circuit shown, the output across the diode will be 10 millivolts per degree Kelvin. Degrees Kelvin are the same as degrees Celsius, except that they're referenced to absolute zero instead of the freezing point of water (note: 0 degrees C = 273 degrees K).

If you merely want to measure the temperature, then install the LM-335 "diode" on the PCB and solder-tack the wires to it. The temperature can then be measured with an ordinary voltmeter. Otherwise, mount it permanently on the PCB. Another application is to use the voltage from the LM-335 to turn on a fan or an alarm when the temperature reaches a certain critical limit. A high-power commercial transmitter uses one of these devices on each PCB and inside each subassembly compartment and then monitors all of them with a multichannel A/D converter connected to a small "single-board computer/controller." A shut-down program can turn off the transmitter in an orderly manner — or warn the operator — when the temperature gets too high.

conclusion

Heat is the great destroyer of electronic components. If a piece of equipment runs too hot, then the result will be unreliable operation, frequent breakdowns, and all the headaches that accompany low reliability. The simple methods shown in this article will enable you to build and/or modify equipment to gain the longest and most reliable use possible.

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VHF/UHF WORLD

Joe Reisert W1JR

33 cm — our newest band

CQ, CQ, CQ, this is KM2XMS calling CQ 33 centimeters . . .

Silence . . . except for white noise.

No one in my area had gear, although a few of us had FCC Experimental Licenses with those funny call signs. But now that's all changed, because on September 28, 1985, the FCC released our newest band, 33 cm, 902-928 MHz. Unfortunately for me, an initial band opening QSO was cancelled as hurricane Gloria flattened our area that very day. Power wasn't restored for five days, so no opening day QSO's were possible. (Yes, I have a generator on order, but so do 5000 other people in my area.)

Now that it's ours, the new 33-cm band holds lots of promise for Amateurs. It's a generous chunk of spectrum — 26 MHz — nestled between the prime real estate of the communications companies and adjacent to the UHF television band. This means that there should be more equipment (components especially) available, than for the 23-cm band (1240-1300 MHz), where high power linear tubes are scarce.

Amateur Service on the 33-cm band is secondary to industrial, scientific, and medical (ISM), but this probably won't cause too many Amateurs any grief. However, restrictions will apply to Amateurs in Colorado, Wyoming, the United States possessions in Region 3 and those hams located near the White Sands Missile Range. The rest of us should enjoy a clean spectrum free of spurious generators and radars.

In order to get the ball rolling on 33 cm, I've updated the material I presented at the Eighth Annual Eastern VHF/UHF Conference in Nashua, New Hampshire, on May 17, 1983. This month's column will illustrate these entry-level circuits and techniques and should provide the necessary impetus to generate activity on the 33-cm band until more Amateur designs and commercial gear are forthcoming.

overview of the band

Our newest Amateur UHF band, large compared with the lower VHF and HF bands, permits the greatest variety of authorized transmission modes. Hence, there's considerable interest in how the band will be subdivided among the various interest groups.

The ARRL VUAC (VHF/UHF Advisory Committee), in conjunction with the VRAC (VHF Repeater Advisory Committee) has set up the interim band plan shown in **table 1.**[1] Note that the narrow-bandwidth, weak-signal segment (the frequencies to which this column is usually dedicated) is the lower 2 MHz of the band. Of prime interest is the weak signal calling frequency, 903.1 MHz, around which most of the communications on CW and SSB will probably prevail.

Radio propagation on this band will be very similar to that experienced on the 70- (420-450) and 23-cm bands. Foliage attenuation will be more of a problem on 33 cm than on 70 cm, but scatter propagation should be better. This band should be perfect for EME, since small (i.e. 12-15 feet or 3.5-4.5 meter) diameter parabolic dishes should be sufficient to produce reasonable echos with 500 watts of transmitted power at the antenna feed. Additional information on propagation can be found in references 2 and 3.

table 1. Abbreviated ARRL 33-cm Interim VUAC Bandplan. See Ref. 1 for more detail.

Segment	Use
902-904 MHz	Narrow-bandwidth weak-signal communications with 903.0-903.05 exclusively for EME and 903.1 MHz as the weak signal calling frequency.
904-906 MHz	Digital communictions.
906-907 MHz	Narrow-bandwidth FM simplex with 906.5 MHz as National simplex frequency.
907-910 MHz	FM repeater inputs.
910-916 MHz	ATV.
916-918 MHz	Digital communications.
918-919 MHz	Narrow-bandwidth FM control links and remote bases.
919-922 MHz	FM repeater outputs.
922-928 MHz	Wide-bandwidth experimental, simplex, ATV, spread spectrum.

antennas and transmission lines

This band is in a transitional antenna region. While Yagi types of antennas should work, they will require close tolerances (0.04 inch or 1 mm) if the desired performance is to be attained.

The most probable antennas will be the parabolic dish and the loop Yagi. Details on parabolic dish design and construction can be found in references 4 and 5. Several loop Yagi designs are described in reference 6.

The loop Yagi designs described in reference 6 can be scaled to the 33-cm band, but they will be either too long or too short, based on commonly available boom material. Therefore, I designed a 12-foot (3.65 meter), 33-element loop Yagi for 903 MHz using standard material stock. Its construction is shown in **fig. 1**.

This loop Yagi design should be duplicated exactly as shown if the gain of 19-19.5 dBi is to be attained. If any changes in the boom diameter, loop thickness or width are desired, the loops must be lengthened or shortened accordingly. This procedure, described in detail in reference 6, should be followed very closely.

Transmission lines must be carefully chosen. RG-8 and RG-213/U types should be used sparingly since they have a loss of about 8 dB per 100 feet (30.5 *meters*). Belden 9913, hardline and Heliax™ are recommended. A thorough discussion of transmission line selection and nominal losses are covered in reference 7.

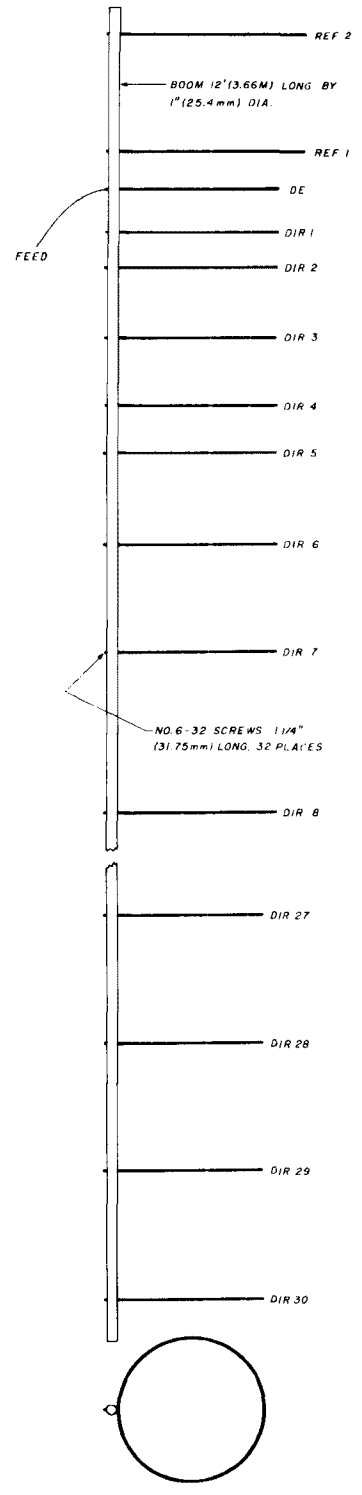
up/down converters and transverters

Receive and transmit up/down converters are often used on the VHF/UHF bands. More recently, transverters have been gaining popularity; the advantages and disadvantages were discussed in references 8 and 9, so they won't be repeated here.

Suffice it to say that transmit upconverters/transverters are preferred to multipliers since they will allow CW and SSB to be used at will. Furthermore, I recommend the modular approach to design, especially since this band is new and components and circuits can be easily upgraded as the available devices are selected and designed into improved circuits.

mixers

When designing a linear up/down



Element	Spacing	Circumference
Refl. 2	1.000	13.949
Refl. 1	5.454	13.949
Dr.El.	6.819	13.313
Dir. 1	8.428	11.900
Dir. 2	9.621	11.900
Dir. 3	12.178	11.900
Dir. 4	14.736	11.900
Dir. 5	16.532	11.900
Dir. 6	19.851	11.900
Dir. 7	24.966	11.900
Dir. 8	30.081	11.900
Dir. 9	35.196	11.900
Dir. 10	40.311	11.900
Dir. 11	45.426	11.900
Dir. 12	50.541	11.539
Dir. 13	55.656	11.539
Dir. 14	60.771	11.539
Dir. 15	65.886	11.539
Dir. 16	71.001	11.539
Dir. 17	76.116	11.539
Dir. 18	81.231	11.107
Dir. 19	86.346	11.107
Dir. 20	91.461	11.107
Dir. 21	96.576	11.107
Dir. 22	101.691	11.107
Dir. 23	106.806	11.107
Dir. 24	111.922	11.107
Dir. 25	117.037	11.107
Dir. 26	122.152	11.107
Dir. 27	127.267	11.107
Dir. 28	132.382	11.107
Dir. 29	137.497	11.107
Dir. 30	142.612	11.107

Notes:
1. Dimensional tolerances should be held to ±0.013 with ±0.04 inch maximum.
2. Reference all spacing from end of boom to prevent tolerance buildup.
3. Loop length should be approximately 1/2 inch longer than circumference shown to allow 1/4-inch overlap on each end of strap. The circumference dimension shown is the actual distance between holes as shown. The straps on this design are made from 0.062-inch thick aluminum 3/8 inch wide. Any deviations from this width or thickness must be compensated for or performance will deteriorate.
4. The driven element is made from brass 0.020 inch thick and 3/8 inch wide. Details are shown below.
5. Adjust height for best VSWR.

fig. 1. A recommended 33 element loop Yagi for the 33 cm band. Gain is approximately 19 dBi and beamwidth is approximately 20 degrees. Recommended stacking distance is 34 inches (86 cm) in the E plane and 32 inches (81 cm) in the H plane. See reference 6 and text for further details.

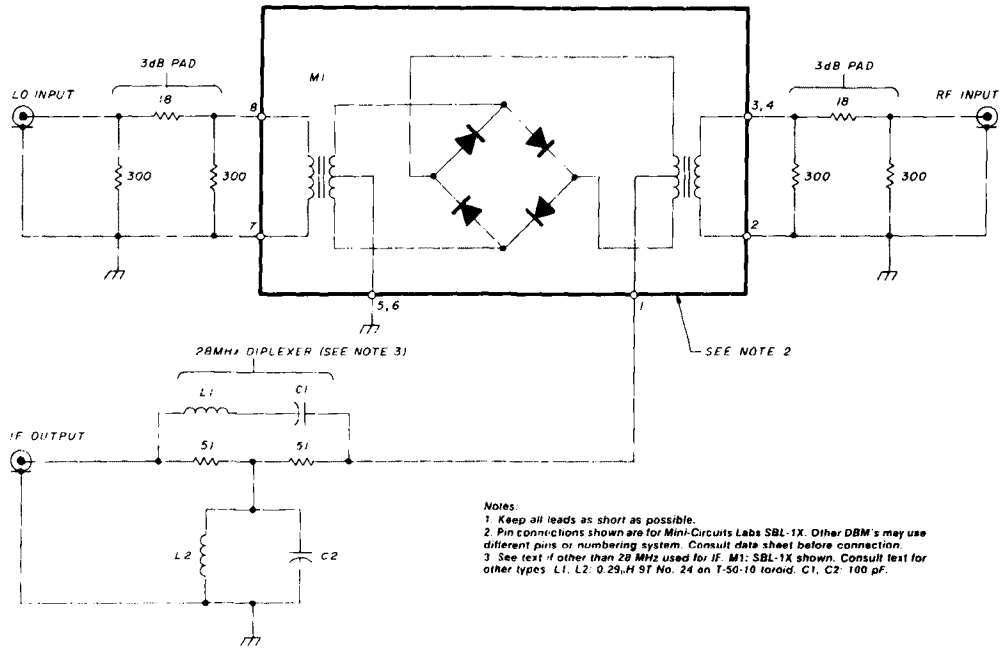


fig. 2. A recommended mixer circuit for a receive type down-converter. Conversion loss is approximately 9 dB overall. LO level should be between 5 and 15 milliwatts. If a 28 MHz IF is not used, the diplexer must be modified as described in text. See text for other recommended dBm's.

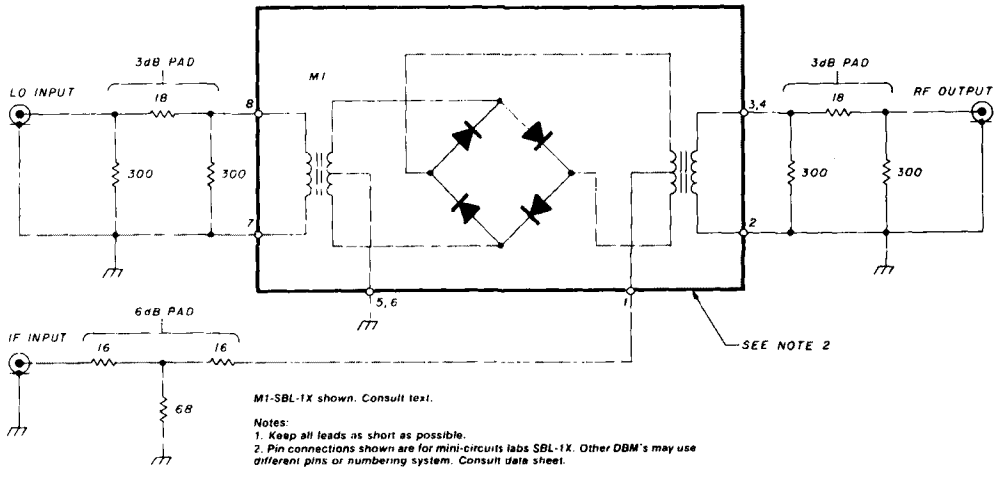
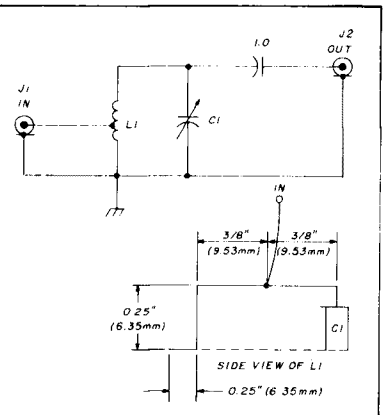


fig. 3. A recommended mixer circuit for a transmit type up-converter. Overall conversion loss is approximately 9 dB. LO level should be between 5 and 15 milliwatts. IF input level should not exceed 1 milliwatt. See text for other recommended dBm's.

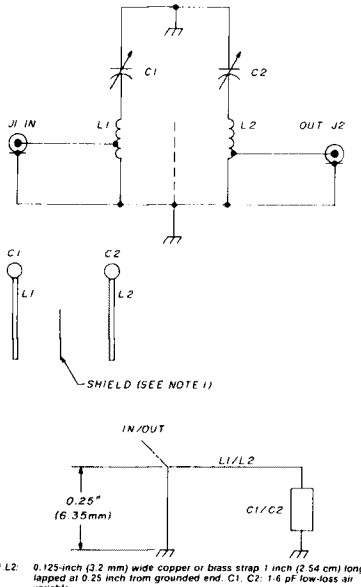
converter, the first requirement is to choose a mixer. For many of the reasons mentioned in references 8 and 9, I recommend the doubly-balanced mixer (DBM) and, more specifically, the commercial packaged units. Many are available, but they must be carefully chosen since most of the commonly available types are restricted to 500 MHz and down.

The Minicircuits Labs SBL-1X (at $5.95), the SBL-1Z (at $6.95), the TFM-2 (at $11.95) — all prices are given for quantities of 10 to 49 — or the Anzac Electronics MD 110 (at



J1,J2: Low loss connectors.
L1: 0.25-inch (6.35 mm) wide brass or copper thin strip 1.25 inch (32 mm) overall length tapped as shown in sketch. C1: 1-6 pF low-loss air variable recommended

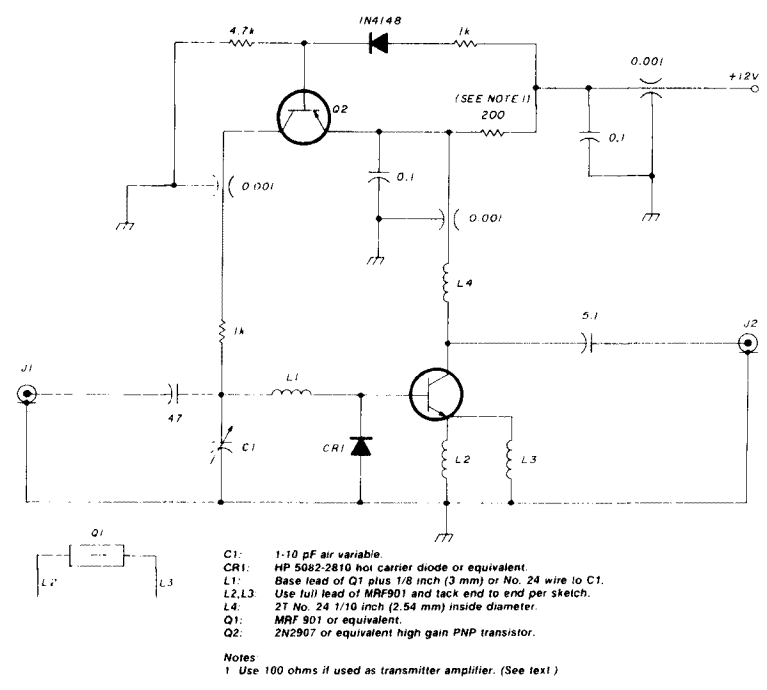
fig. 4. A simple single section 903 MHz bandpass filter for protecting the front end of a receiver. Bandwidth is approximately 100 MHz and insertion loss is 0.5 dB maximum.



L1,L2: 0.125-inch (3.2 mm) wide copper or brass strap 1 inch (2.54 cm) long tapped at 0.25 inch from grounded end. C1, C2: 1-6 pF low-loss air variable.

Notes:
1. L1 and L2 are in parallel and spaced 0.5 inch (1.27 cm) center to center. The shield is 0.5 inch (1.27 cm) long and placed in middle between L1 and L2. It extends from the grounded end upwards per sketch.

fig. 5. Circuit of a simple two-section bandpass filter suitable for a 33 cm low-level receiver or transmitter. Bandwidth is approximately 50 MHz and insertion loss is 0.5 dB.



C1: 1-10 pF air variable.
CR1: HP 5082-2810 hot carrier diode or equivalent.
L1: Base lead of Q1 plus 1/8 inch (3 mm) or No. 24 wire to C1.
L2,L3: Use full lead of MRF901 and tack end to end per sketch.
L4: 2T No. 24 1/10 inch (2.54 mm) inside diameter.
Q1: MRF 901 or equivalent.
Q2: 2N2907 or equivalent high gain PNP transistor.

Notes
1 Use 100 ohms if used as transmitter amplifier. (See text)

fig. 6. A simple low noise preamplifier for a 33 cm receive type converter. Gain is approximately 12 dB and noise figure is 2.5-3 dB typical. If the 200 ohm resistor in the constant current source is changed to 100 ohms, it will function well as a low-level linear transmitter amplifier stage as discussed in text.

$19.00 each) are recommended. DBMs from other suppliers are likewise usable as long as they're specified to work up to at least 1 GHz.

A recommended receive-type down-converter DBM circuit is shown in **fig. 2.** I prefer 28 MHz for an IF. The diplexer shown on this circuit is for 28 MHz, per reference 9. Other IF's can be used, but the diplexer shown will have to be scaled to the new IF frequency or be eliminated. A recommended IF post amplifier is described in reference 10.

Figure 3 shows a recommended low-level DBM circuit for a transmit up-converter. The operation of this circuit is described in reference 8. This circuit will easily handle any desired IF up to 150 MHz. The IF input level must not exceed 1 milliwatt.

filters

Figure 4 shows a simple input filter that can be used ahead of a receive down-converter, especially if the input stage is untuned. It is not exotic, but will eliminate such out-of-band signals as TV, FM, etc. This filter is easier to build than a coaxial cavity, and its unique topology has a symmetrical response.[11]

A simple two-section bandpass filter is shown in **fig. 5.** It should be used in the receive down-converter just ahead of the mixer to eliminate any out-of-band signals from reaching the mixer. This filter should also be used after the transmit mixer to prevent amplification of local oscillator, image and spurious signals generated by the mixer from being amplified in the transmitter.

These filters are simple, but neither is real state-of-the-art. Interdigital types of filters with three sections are recommended for improved filtering per reference 11, but are beyond the scope of this month's column.

low-level receiver preamplifiers.

The MRF 901 bipolar transistor is a readily available (Radio Shack), low-cost device (under $2.00) that is simple and straightforward to use. A

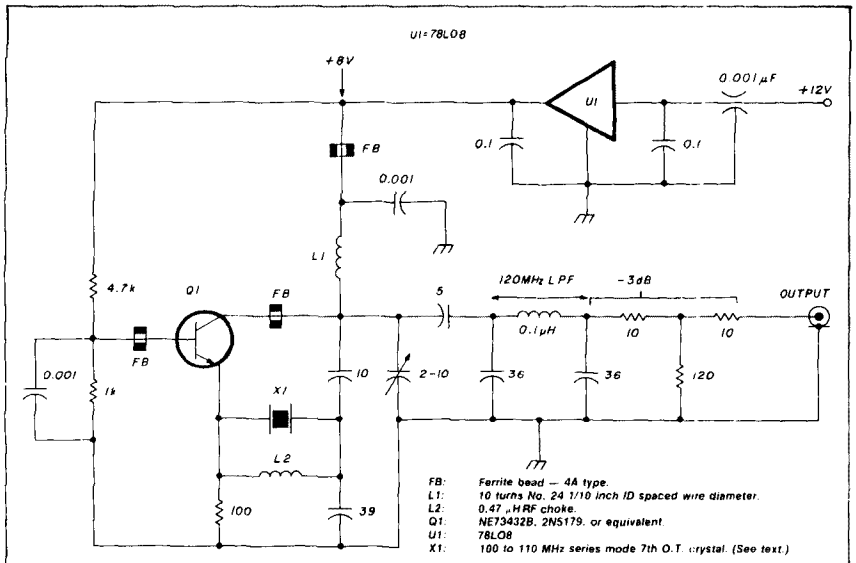


fig. 7. A recommended local oscillator circuit for a 33 cm up or down-converter. Any frequency between 100 and 115 MHz will work well. 109.3625 MHz is recommended for a 28.1 MHz IF. Output power level is approximately 10 milliwatts.

recommended circuit patterned after a previous design[12] is shown in **fig. 6**. It uses series feedback and simple matching to achieve a moderate 2.5-3 dB noise figure. It also has some built-in selectivity and reasonably high output power (over 5 milliwatts at 1 dB compression).

A single preamplifier stage such as this one will normally be sufficient to yield an overall 3-4 dB noise figure in a typical converter as just described. Two such preamplifers can be used in cascade if a lower noise figure is desired.

If a very low noise figure (less than 1 dB) is required, the 902 MHz GaAs FET preamplifier in reference 13 can be used. However, this particular design has little if any input selectivity to reject transient or lower frequency emitters. Therefore, if this circuit is used, I recommend adding a 25-100 pF, low-loss ceramic chip-type capacitor in series between the input connector and the first circuit elements.

Other GaAs FET circuit recommendations are described in reference 14. Reducing the size of the input inductor and capacitors in the circuit in this reference should yield a very acceptable noise figure with "built-in" front-end selectivity, thus killing two-birds with one stone.

local oscillators and multipliers

So far I have not mentioned a suitable local oscillator. The 33-cm band is unique in that it can be easily served with a simple crystal oscillator operating in the 100 MHz region and followed by three doublers. This is a recommended approach.[9]

Figure 7 shows a recommended oscillator circuit similar to the one described in reference 9. If a crystal cut for 109.3625 MHz is used, the IF for 903.0 MHz will be 28.1 MHz, a favorite IF of mine.[9] This circuit has been widely used. A low-pass filter has been added to the output to decrease oscillator harmonics. I recommend placing this oscillator in its own shielded box, away from heat and extraneous RF signals.

Figure 8 shows a recommended multiplier circuit that consists of three doublers. It has a clean output and is relatively easy to align. The RF output level is sufficient to directly drive the DBM circuits. This circuitry is similar to that described in reference 9 and has been extended to the 33-cm band. It should also be placed in its entirety in a shielded box.

If a transverter is used, the multiplier output power is sufficiently high enough so that it can be divided into two equal outputs. A Wilkinson-type power splitter is recommended since it has negligible loss (over the inherent 3 dB power split) and provides high isolation (20 dB typical) between the two outputs. Hence there will be very little, if any, interaction between the receiver and transmitter. The Wilkinson-type power splitter I use is shown in **fig. 9**. Both transmission lines are 75 ohms and are electrically a quarter-wavelength long at the local oscillator output frequency.

transmitter circuits

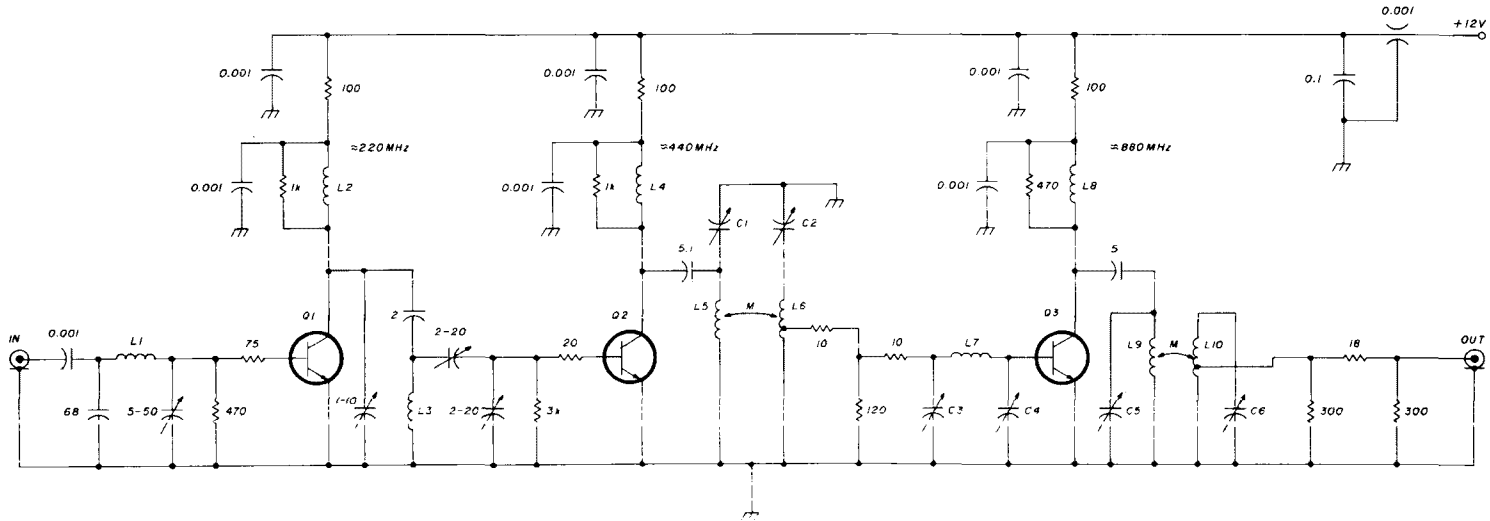
Finally we come to the transmitter. The output of the DBM shown in **fig. 3** will be about 16-17 dB below a milliwatt with one milliwatt of IF drive, the maximum recommended level for a clean transmitter output. The DBM should be followed by either the filter shown in **fig. 5** or an equivalent as described.

This low-level output after the filter can be easily boosted up to a moderate power level with two amplifier stages, similar to the receive preamplifier shown in **fig. 6**. All that's required is to change the 200-ohm resistor in the constant current source to 100 ohms and remove the protection diode, CR1. Gain will then be about 13 dB per stage and the 1 dB output compression point will increase to about 10 milliwatts.

Alternatively, Toshiba and NEC now make low-cost ($7-10), 902-905 MHz low-level linear hybrid modules. The Toshiba module part number is S-AU15; the NEC model part number is MC-5809.* Both units require about 8 volts DC. Gain is just over 20 dB and the 1 dB output compression point is

*Toshiba is represented in the United States by Matcom, Inc., 450 San Antonio Road, Palo Alto, California 94036.

**NEC is represented in the United States by California Eastern Labs, 3260 Jay Street, Santa Clara, California 95054.



C1-C6: 1-10 pF air variables.
L1: 12T No. 24 C.W. 1/10-inch ID
L2,L3: 4T No. 24 1/4-inch long 1/10-inch ID.
L4: 0.1μH RF choke.
L5,L6: 1/8-inch brass strap, 3/4-inch overall length, spaced 3/4 inch L6 tapped 1/4 inch above ground.
L7: 4 turns No. 24 1/10-inch ID, 1/4 inch long.
L8: 8 turns No. 24 1/10-inch ID, 1/4 inch long.
L9,L10: Same as L5, L6. Tap 1/4 inch up on L10.
Q1,Q2: NE73432B or 2N5179.
Q3: NE02132 preferred, but NE73432B OK at slightly lower power output.

fig. 8. A recommended multiplier circuit suitable for a 33 cm local oscillator. Each stage is a doubler with the final output frequency in the 800-900 MHz region. The oscillator in figure 7 is recommended as the driver. Output power is approximately 10-20 milliwatts with an input of 5-10 milliwatts.

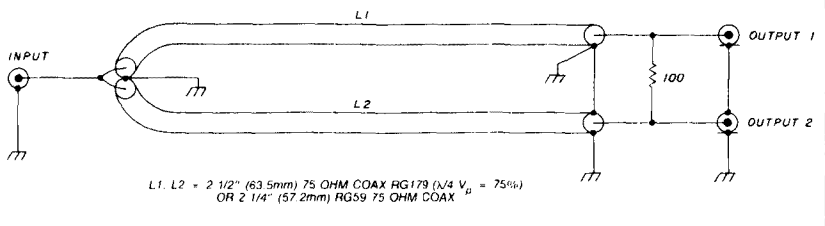


fig. 9. A Wilkinson type two-way in-phase power splitter suitable for local oscillator power splitting in the range of 800-900 MHz. Other frequencies can be used by adjusting the length of the 75 ohm coax for 0.25 electrical wavelength at the desired frequency.

over 100 milliwatts. A typical circuit using these modules is shown in **fig. 10.**

For higher linear power, CATV-UHF type bipolar transistors can be used. A recommended circuit, patterned after the circuits described in reference 8, is shown in **fig. 11**. Gain is typcially 13 dB per stage with a 1 dB compression point of 300 milliwatts. One or two stages can be used, depending on the desired gain and output power. This power level is more than adequate for local (i.e. up to 25 miles or 40 km) QSO's.

For even higher solid-state linear output power, transistors similar to the NEC NE0801 (11 watts) are recommended.[2] I'm sure there are many other devices available from suppliers such as Acrian, Motorola, TRW, and Thompson-CSF's Solid State Microwave Division. Time and space does not allow for a detailed description of such circuitry at this time.

For "quick and dirty" gain, Class "C" can be used. The same suppliers just mentioned can supply suitable class "C" bipolar transistors to at least 25-50 watts.

Furthermore, if only class "C" operation is desired, both NEC and Toshiba make 7-12 watt output hybrid modules for the 33-cm band, which is, incidentally, a citizens' band in Japan. The NEC part number is MC-5843 and the Toshiba part number is S-AU11.

A suitable circuit using these modules is shown in **fig. 12**. These hybrid modules provide a power gain of approximately 30 dB and can be driven

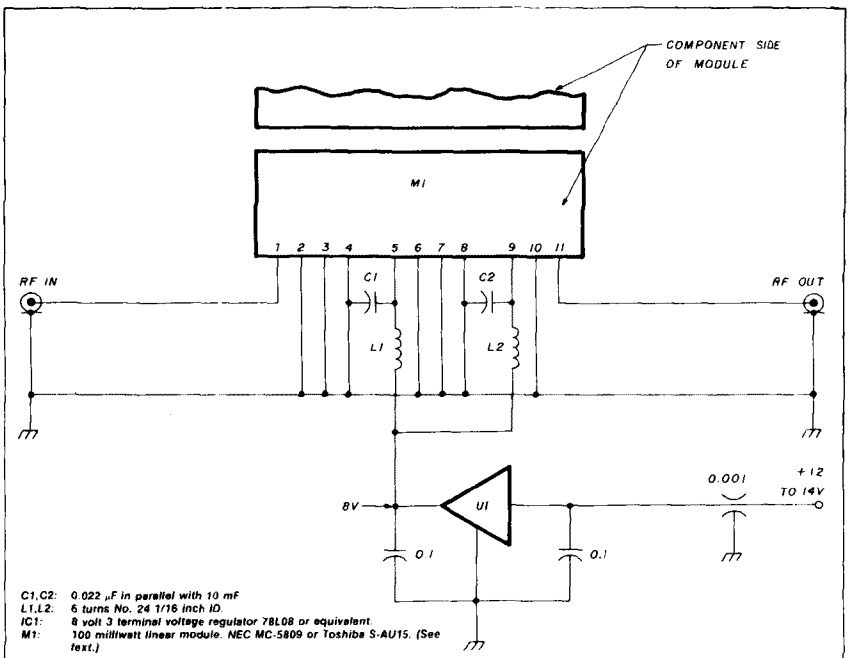


fig. 10. A recommended low-level linear transmit amplifier circuit using commercial hybrids. Unit requires approximately 8 volts at 100 milliamps. See text for further details and suppliers.

to full output with 100-200 milliwatts of drive. They require a nominal supply of 12.5 volts at 2-3 Amperes of current and are great for portable operation.

high power

High-power amplifier designs are probably already available, but we have to seek them out. For moderate power (25-200 watts), the ubiquitous 2C39/7829 in a cavity is recommended.[14,15] I'm sure Amateur designs using these tubes will be published shortly.

For even higher power, I'm aware of only one published Amateur design.[16] UHF TV tansmitting tubes should be readily available, especially as "pull-outs." The RCA 7650 and 7213 immediately comes to mind. Cavity-type amplifiers using these or other suitable tubes are recommended.[15,16] I'm sure that many designs will be forthcoming as interest picks up in this new band.

summary

This month's column was mainly focused on getting started on the new 33-cm band. Easy-to-build and duplicate circuitry was discussed. Although the power level available from this is low, it should be more than adequate for DX from 50 to 250 miles (80 to 400 km) for band "warmer-upper's" and further if extended propagation conditions are present.

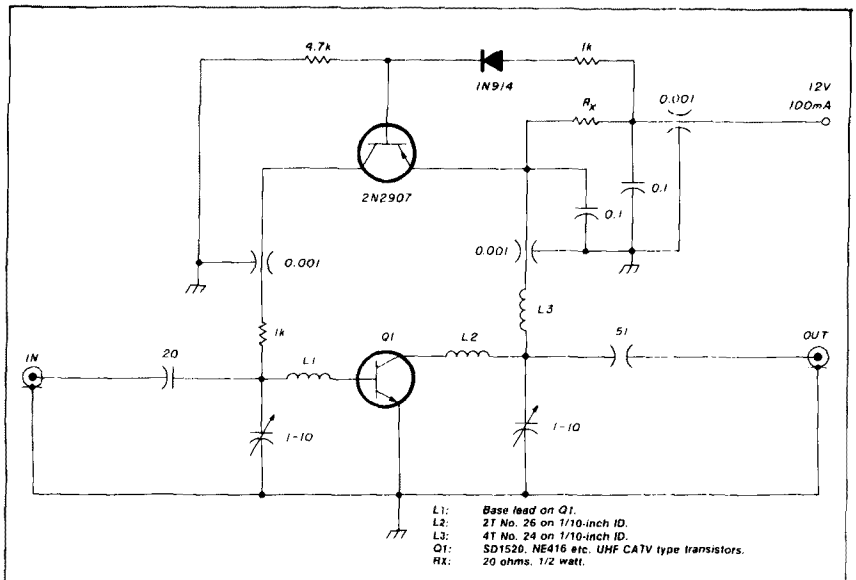


fig. 11. A recommended medium-power linear transmit amplifier. Gain is approximately 13 dB and maximum output power at 1 dB compression is approximately 300 milliwatts.

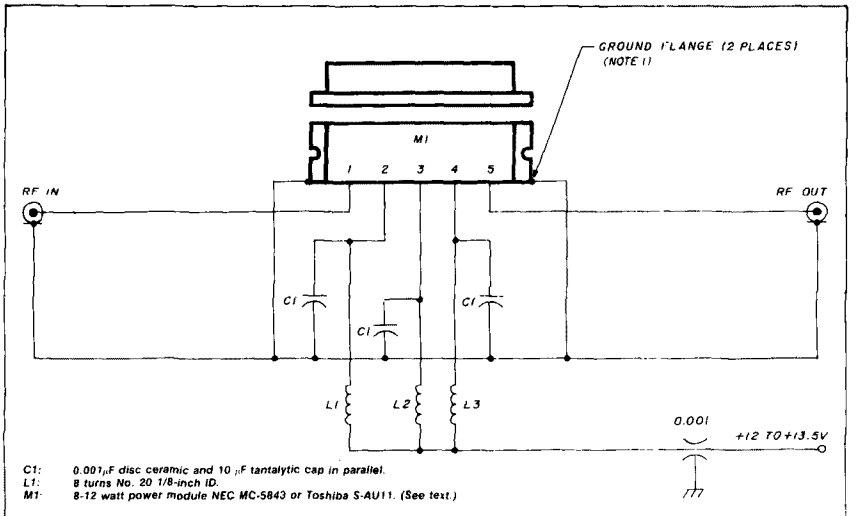


fig. 12. A recommended medium power class "C" amplifier circuit using commercial hybrids. Gain is approximately 30 dB and output is 7-12 watts with 100 milliwatts drive. See text for suppliers. Be sure to provide adequate heat sink with several radiating fins. Under normal operation the heat sink will have to dissipate up to 25 watts of power.

The designs just discussed are more than adequate for transmitter drivers and basic receive converters. Improved designs and higher power transmitter designs should be forthcoming and can be easily substituted or added on to the circuitry shown, especially if the modular approach is used.

Let's welcome our "newest" UHF band. It was only a few years ago that power levels above 10 watts were uncommon on 23 cm, and we all know that great DX was worked there under good conditions. I'm sure the same DX is more probable on 33 cm. See you on 903.1 MHz!

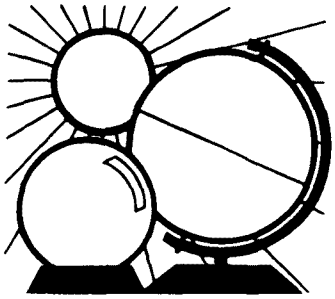
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12. Bill Smith, K0CER, "The World Above 50 Mc.— W6FZJ Wide-Band Low-Noise Preamplifier," *QST*, November, 1972, page 112.
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16. Joe Reisert, W1JR, "VHF/UHF World: High Power Amplifiers, Part 2," *ham radio*, February, 1985, page 38.
17. Robert I. Sutherland, W6PO, and William I. Orr, W6SAI, "A High-Power Cavity Amplifier for the New 900 MHZ Band," *QST*, November, 1982, page 14.

important VHF/UHF events:

April 14: *ARRL 144-MHz Sprint Contest*
April 19/20: *REF IARU EME Contest*
April 21: *Predicted peak of Lyrids meteor shower at 1945 UTC*
April 22: *ARRL 220-MHz Sprint Contest*
April 25: *EME perigee*
April 25-27: *Dayton Hamvention*
April 30: *ARRL 432-MHz Sprint Contest*
May 4: *Predicted peak of the Eta Aquarids meteor shower at 1900 UTC*
May 8: *ARRL 1296-MHz Sprint Contest*
May 10/11: *So. Calif. 6 Meter Club QSO Party (contact N6FSL)*
May 16/18: *12th Annual Eastern VHF/UHF Conference, Nashua, NH (contact W1EJ)*
May 17: *ARRL 50-MHz Sprint Contest*
May 24: *EME perigee*

ham radio



DX FORECASTER

Garth Stonehocker, KØRYW

1985 review

Six months have passed since our last review of propagation conditions; now that data for 1985 is complete, it's time for a review of the year as a whole.

The sunspot numbers (SSN) during the first four months of 1985 were just under 20, increasing through July to 35 and diminishing during the final months of the year to 13. This represented a decrease, over the year, of 7 SSNs, or approximately 0.5 SSN per month. If this trend continues, we could expect a low of about 6 SSNs by late fall of 1986. Although new 11-year cycle SSN sunspots were tentatively identified from polarization differences on September 11, no opposite polarity spots at high latitudes have yet been seen. The cycle 21 SSN minimum may occur in September, 1986.

An equivalent pattern emerged in review of the 10 cm solar flux data. *Monthly* solar flux numbers in early 1985 matched those of late 1984 at 74 ±2 units. A mid-year bulge of up to 80 occurred instead of the minimum number expected; a minimum of 69.5 occurred in September. The year closed with a 75 again. Note that the minimum monthly average in October, 1984, was 73.5; in 1985, it was 69.5, which represents a decrease of four units. A continued reduction to 67 units in a late summer month of 1986 will be necessary for the solar flux minimum to be reached.

The lowest *daily* solar flux value in the 11-year cycle so far, 66, occurred on August 17 and from October 6 through the 9th. The highest value during 1985 was 101, on July 9. (The daily flux number will probably reach 63 or 64 by next fall.) The 27-day solar flux variation of daily numbers, whose increase raise the monthly average, was 20 units or more in January, in the period between April and July (with May the highest solar flux, at 80.3), and in the month of October; a variation of about 10 units occurred in February, March,, August, November, and December. September was quite "flat," devoid of any 27-day solar cycle activity, and therefore provided the minimum monthly average of the year. Four days of a solar flux of 66 were the year's minimum recorded just before solar activity began to increase again in mid-October.

The geomagnetic *A* figure monthly average was, as usual, highest in April, the most disturbed month, and May, the quietest. Note the May inverse relationship to solar flux, a surprisingly frequent occurrence. Of the disturbed periods, April 20 to May 2 stands out as the highest and longest. The other months had two or three milder periods of disturbance separated by a few days of quiet conditions. That's the difference between the geomagnetic conditions in the winter minimum months and those of the equinox maximum and summer months: the number of big events decreases and the number of quiet days (at a lower *A* level) increases. As SSN minimum approaches this summer, the periods of solar 27-day activity should get further apart and be lower in intensity. The periods of geomagnetic disturbance should also decrease in number and intensity. The sun simply takes a rest — Hi!

You may recall from my December, 1985, column on maximum usable frequencies (MUF) that an increase of one unit in solar flux results in an increase in MUF of 1 percent. The base line to start from for mid-latitude MUFs is MUF=2.5 (0.036 SSN + 5.28). The decrease in MUF from a geomagnetic disturbance is percentage MUF= 0.375A + 3.75. Listen to WWV at 18 minutes after the hour for the data.

last-minute forecast

The first two weeks of April are expected to have low solar flux levels. Because the amplitude variation isn't more than a few units during this part of the 11-year sunspot cycle, the lower frequency bands are expected to provide the best DX activity. The geomagnetic field should still be a problem this close to equinox, so look for disturbances around the 4th to the 8th, from the 18th to the 22nd, and again on the 28th of the month. Signal level variations will be greatest during these periods and conditions won't be really stable in between, either, since equinoctial periods don't produce many quiet days. The best days for the higher frequency band DX are more likely to occur from the end of the second week through the third. Expect some enhanced equatorial openings during the disturbances of the 21st.

The perigee of the moon's orbit (for moonbounce DX) is on the 25th, with the moon showing full phase on the 1st. There will be a short meteor shower, the Lyrid, on April 20-22, with a rate of five per hour — hardly much help for meteor-scatter DX. But a bigger shower, the Aquarid, starts before the end of April, peaks on May 5, and ends in mid-May. Its rate is 10 to 30 per hour.

band-by-band summary

Ten, twelve, fifteen, and twenty meters will be open from morning until early evening almost every day, and to most areas of the world. The openings on the higher of these bands will be shorter to the southern hemisphere and will occur closer to local noon. Transequatorial propagation on these bands will more likely occur toward evening during conditions of highest

GMT	PST	WESTERN USA N	NE	E	SE	S	SW	W	NW	MST	MID USA N	NE	E	SE	S	SW	W	NW	CST	EST	EASTERN USA N	NE	E	SE	S	SW	W	NW
0000	8:00	20	30	20	10	15	10	10	15	5:00	20	30	20	10	15	10	10	15	6:00	7:00	20	30	20	10	15	10	10	20
0100	5:00	20*	*40*	20	12	15	10	10	15	6:00	20	30	20	12	15	10	10	15	7:00	6:00	20	30	20	12	15	10	10	20
0200	0:00	20	*40*	*20*	12	15	10	10	15	7:00	20	*40*	*20*	12	20	10	10	20	8:00	9:00	30	40	20	15	20	10	10	20
0300	7:00	20	*40*	*20*	15	15	10	10	15	8:00	30	*40*	*20*	15	20	10	10	20	9:00	10:00	30	*40*	*20*	15	20	12	10	30
0400	8:00	20	30	*20*	20	20	10	10	20	9:00	30	*40*	*20*	20	20	12	12	20	10:00	11:00	*30*	*40*	*20*	20	30	12	12	30
0500	9:00	20	30	*20*	20	20	12	10	20	10:00	30	30	*20*	20	30	12	15	20	11:00	12:00	*40*	*40*	20	20	30	15	15	30
0600	10:00	20	30	20	20	30	12	12	20	11:00	*40*	30	20	20	30	15	15	30	12:00	1:00	*40*	*40*	20	20	30	15	15	*30*
0700	11:00	30	*30*	20	20	30	15	12	20	12:00	*40*	*40*	20	*20*	30	15	15	30	1:00	2:00	*40*	*40*	*20*	*20*	30	20	20	*40*
0800	12:00	30	*40*	20	*20*	30	20	15	30	1:00	*40*	*40*	*20*	*20*	30	20	20	*30*	2:00	3:00	20	30	*20*	*20*	30	20	20	*40*
0900	1:00	*30*	*40*	*20*	*20*	30	20	20	30	2:00	30	20	*20*	*20*	30	20	20	*40*	3:00	4:00	20	20	20	*20*	40	20	20	*40*
1000	2:00	*40*	*40*	*20*	*20*	30	20	20	30	3:00	20	20	20	*20*	40	20	20	*40*	4:00	5:00	20	20	15	*20*	*40*	*20*	*20*	20
1100	3:00	*40*	20	20	*20*	*30*	20	20	*30*	4:00	20	20	15	*20*	*40*	20	20	*40*	5:00	6:00	20	20	12	20	*40*	*20*	*20*	20
1200	4:00	30	20	15	*20*	*30*	20	20	*40**	5:00	20	20	12	20	*40*	20	20	30	6:00	7:00	20*	20	12	15	*40*	20	20	30
1300	5:00	20	20	12	20	*30*	20	20	*40*	6:00	20	15	10	15	*40*	20	20	30	7:00	8:00	20	20	10	12	*40*	20	20	30
1400	6:00	20	20	10	15	*30*	20	20	*30*	7:00	20	15	10	12	*40*	20	20	30	8:00	9:00	20	15	10	12	*40*	20	20	30
1500	7:00	30	15	10	12	*30*	20	20	*30*	8:00	20	15	10	12	20	20	20	*30*	9:00	10:00	20	15	10	12*	20	20	20	*30*
1600	8:00	30	15	10	12	20	*20*	20	*30*	9:00	30	15	10	10	20	20	20	*40*	10:00	11:00	30	20*	10	10	20	20	20	*40*
1700	9:00	*30*	15	10	10	20	20	20	*40*	10:00	30	15	10	10	20	20	*30*	*40*	11:00	12:00	30	20	10	10	20	20	*30*	*40*
1800	10:00	*40*	20*	10	10	15	15	*30*	30	11:00	30	20*	10	10	15	15	20	30	12:00	1:00	*30*	20	10	10	15	15	*30*	30
1900	11:00	*40*	20	10	10	15	15	20	20	12:00	*40*	20	10	10	15	12	15	20	1:00	2:00	*30*	20	10	10	15	12	15	20
2000	12:00	*40*	20	12	10	15	12	12	20	1:00	*40*	20	12	10	15	12	12	20	2:00	3:00	*40*	20	12	10	15	12	12	20
2100	1:00	30	20	12	10	15	12	12	20	2:00	*40*	20	15	10	15	10	12	20	3:00	4:00	*40*	20	15	10	15	10	10	20
2200	2:00	20	20	15	10	12	10	10	15	3:00	20	20	15	10	15	10	10	15	4:00	5:00	30	20	15	10	15	10	10	15
2300	3:00	20	30	20	10	12	10	10	15	4:00	20	30	20	10	15	10	10	15	5:00	6:00	20	30	20	10	15	10	10	15
APRIL	ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN			CARIBBEAN S. AMERICA								

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours

*Look at next higher band for possible openings.

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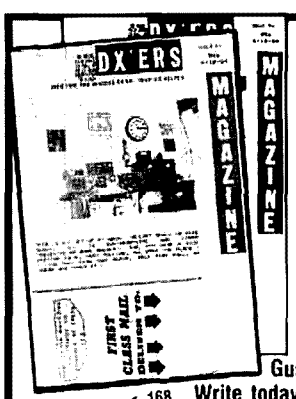
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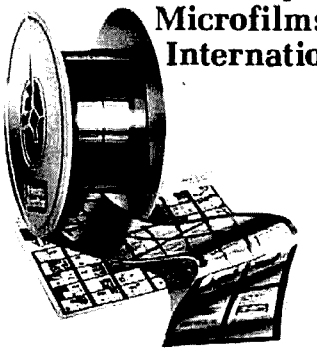
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improved repeater/ transmitter noise performance

Solid-state technology has dramatically improved the reliability of most Amateur repeater systems. However, the changeover from tubes to semiconductors has not always progressed as smoothly as expected for some repeater operators.

Repeater desense is normally caused by inadequate transmitter-to-receiver isolation, allowing gain compression to occur in the repeater receiver. The 100 dB isolation characteristics provided by modern duplexers, when properly installed, make this an unlikely possibility. Alas, the duplexer is often blamed for poor repeater performance even though the problem is actually in the transmitter.

If repeater desense occurs when a new solid-state exciter replaces an older tube type, chances are the problem is caused by excessive transmitter noise products. The standard cure recommended for this is usually the addition of another transmitter notch cavity. Unfortunately, this is an expensive, bulky, and lossy (about 1 dB) solution. Earlier tube exciters, with their hi-Q interstage coupling networks, minimized this problem.[1] The low-Q interstage networks in solid-state transmitters increases the bandwidth and thus the noise output on the receiver frequency.

what can be done?

Commercial repeater manufacturers have been aware that the single stage most often responsible for transmitter noise is the phase modulator. For this reason, most commercial repeater systems now use direct FM techniques — i.e., varactors — rather than phase modulators. Another approach involves the use of a two- or four-pole monolithic filter after the last multiplier stage, or frequency synthesizer, preceding the power stages.[2] (These filters, normally used to increase the IMD performance of the VHF receivers, are capable of handling only about 5 mW of power.)

One advantage of using a crystal filter lies in reducing transmitter noise at all frequencies on either side of the transmitter carrier; in the same application, a notch filter would only protect the receiver channel. Crystal filters are expensive ($100-$200), and considering their power restraints, 50-ohm *termination impedances and insertion* losses might prohibit their use in most existing transmitter designs.

Repeater operators seeking a new solid-state exciter should look for one using low-noise direct FM. Avoid phase modulation. If a homebrew transmitter design is being contemplated, consider including both direct FM and a four-pole monolithic filter to further reduce the transmitter's noise products. The improvements are dramatic and well worth the effort.

references

1. *FM and Repeaters for the Radio Amateur*, First Edition, American Radio Relay League, 1972, Chapter 9, pages 148-150.
2. "DigiCap VHF Transmitters," Bulletin No. 100-0006-008, June, 1981, Quintron Corporation, Quincy, Illinois 62301.

Peter Bertini, K1ZJH

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ALPHA DELTA
Tech Notes

ALPHA DELTA ANTENNA and AC LINE PROTECTORS — the inside story

- **Who Needs Them**
- **Do They Really Work**
- **Why Are There Several Different Models**

Who Needs Them

Lightning is the most common cause of component damage. However, we occasionally run into those who say "I've never been hit by lightning" or "I live on the West Coast and we don't have much lightning." Don't be fooled. There are demons lurking everywhere from your AC line to antenna that can damage your gear. Before exposing those, let's look at data about thunderstorms.

On average, the number of annual days with thunderstorms per area are approximately: West Coast, 5; Southwest, 20 to 40; Texas, 40 to 70; Midwest, 40 to 50; East Coast, 30 to 50; South, 50 to 70; and Florida, up to 100! Really, no matter where you live, you should be aware and protected from the potential for lightning-induced damage.

Now, what about what you can't see that *does* damage equipment? Dry desert winds in the Southwest and West Coast, wind driven snow and summer cloud buildup are all generators of enormous amounts of static electricity. Static-induced voltages from any one of these conditions can build up levels of 3 kV or more! If you've ever had the occasion to watch the static discharge jumping from the end of a long wire hanging near a chassis, you'll know what we mean.

What's worse, this type of damage is not always catastrophic. Semiconductors can suffer junction damage and will degrade over a period of weeks or months, causing subtle system problems and a gradual loss of sensitivity.

In the case of AC line protection, semiconductors are known to be damaged by transients caused by AC motors starting and switches, surges from power company "brown-outs" and poor regulation and even the effects of fluoresent lighting. If you have had the chance to see a graphic printout from an AC wall socket analyzer, you wouldn't plug *anything* in again that was unprotected.

So who needs Alpha Delta? Everyone. Regardless of season or geographic location.

a new class of directive antennas

Improve Yagi performance with curved 1.5λ elements

In the May, 1983 issue of *Transactions on Antennas and Propagation*,[1] Chang and Cheng introduced a new class of antennas that appear to offer much promise for VHF use. Based on concepts developed earlier by F. M. Landstorfer,[2] these antennas feature curved elements, each longer than a wavelength and shaped to compensate for the reversals in phase that occur each half wavelength along an element.

With the 1.5 wavelength elements in the classic reflector-driven-director configuration used in the original experiments Landstorfer claimed gains of 11.5 dBi. The same gain in a conventional Yagi using straight half-wave elements would require about nine elements and a much longer boom. While the new design requires greater width, the combination of gain, short boom length, and mounting simplicity form the attractive features of the design.

principle of operation

The general concept and plan of these antennas is shown in **fig. 1**. The center part of the elements resembles a V radiator. The phase center of the V radiation lies along the center axis, and some distance from the apex of the V. A wave radiated from this section will arrive at the other element parts after a time delay that corresponds to a phase rotation. As a result, even though the outer sections are out of phase with respect to the center section, the delayed wave will be at least partly in phase with the waves radiated by the outer sections. This addition of wave components accounts for the increase in gain over a conventional straight-element Yagi.

The design problem presented by these antennas is to determine the shaping of the elements for maximum gain. This subject was addressed by Chang and Cheng in their article.[1] They approximate the current distributions on the array elements by the method of moments, dividing each element into 22 sections and analytically determining the shaping for maximum gain. The computations are extensive, involving a 63 by 63 complex matrix manipulation (a solution requires

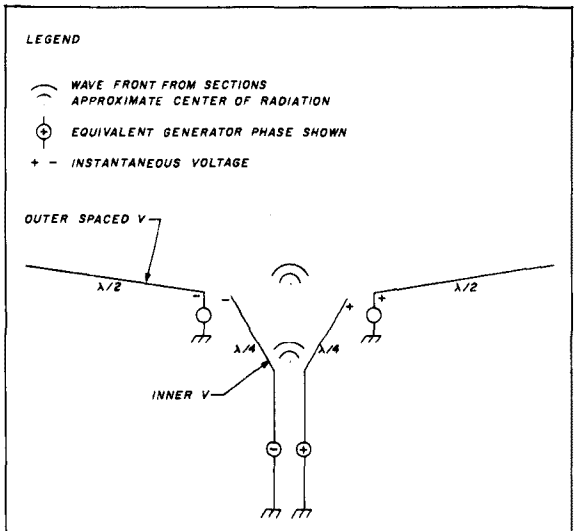


fig. 1. General principle of operation using a shaped element. The center of radiation of the outer half-wave sections is in front of the radiation center for the *center* section, giving both a spacing and a phase difference, as in the ZL-Special. The outer sections may be separately excited, as shown, or joined to the center part. This illustration assumes that the middle half-wave section is center-fed.

By R. P. Haviland, W4MB, 1035 Green Acres Circle, No., Daytona Beach, Florida 32019

```
READY.


10 REM HIGH GAIN YAGI "HGAINYAGI", IN COMMODORE SIMONS BASIC
20 REM R P HAVILAND,6 JUNE 1984
30 REM REFER TO TRANS IEEE AP-31,MAY 1983
40 PRINT "{CLR}"
50 DIM A(3),B(3),C(3),D(3),Z(3)
60 DIM Y(20,4)
70 A(1)=.38:A(2)=.395:A(3)=.364
80 B(1)=20.77:B(2)=53.014:B(3)=204.532
90 C(1)=-.162:C(2)=0:C(3)=.151
100 D(1)=.645:D(2)=.59:D(3)=.55
110 INPUT "PRINTOUT,Y OR N  ";T$
120 PRINT"PRESS SPACE TO END SCREEN DISPLAY"
130 PRINT "ENTER FREQUENCY, MHZ ";
140 INPUT F: LAMDA= 984/F
150 PRINT F: PRINT"ELEMENT LENGTH= ";1.5*LAMDA;" FEET"
160 PRINT "ELEMENT DIAMETER ";12*LAMDA/100;" INCHES"
170 FOR N=1 TO 20: FOR M= 3 TO 1 STEP -1
180 X=(N-1)/20
190 IF X>D(M) THEN Y(N,M)=0:GOTO 220
200 Y(N,M)=A(M)*(1-1/(1+B(M)*X*X))+C(M)
210 Y(N,4)=X
220 NEXTM:NEXTN
230 PRINT"Y VS X COORDINATES, INCHES"
240 PRINT"X";TAB(10);"DIR";TAB(20);"ANT";TAB(30);"REF"
250 FOR N=1 TO 20
260 X=INT (1200*LAMDA*Y(N,4))/100
270 FOR M=1 TO 3
280 Z(M)=INT (1200*LAMDA*Y(N,M))/100
290 NEXT M
300 PRINT X;TAB(10);Z(1);TAB(20);Z(2);TAB(30)Z(3),
310 NEXT N
320 WAIT 197,32
330 IF T$="Y" THEN HRDCPY
340 HIRES 0,1
350 FOR M=1 TO 3:FOR N=2 TO 16
360 T=N-1
370 IF Y(N,M)=0 GOTO 400
380 LINE 160-200*Y(T,4),200*Y(T,M)+40,160-200*Y(N,4),200*Y(N,M)+40,1
390 LINE 160+200*Y(T,4),200*Y(T,M)+40,160+200*Y(N,4),200*Y(N,M)+40,1
400 NEXT N:NEXT M
410 IF T$="Y" THEN COPY
420 WAIT 197,32:NRM

READY.
```

fig. 2. Computer program provides dimensions of a three-element version of the shaped element array. While the program is written in Simons' BASIC for the Commodore 64, translation for other computers should not be difficult.

approximately 40 minutes of DEC-10 computer time). The problem is far beyond the capability of home computers.

program listing aids 3-element design

Fortunately, Chang and Cheng have summarized their results in such a form that makes it possible to duplicate their optimized design for a three-element Yagi array. For convenience, the results have been arranged as a computer program, **fig. 2**, written in Simon's BASIC for the Commodore 64. The program is written for easy translation to other versions of BASIC; only the graphic generation section may require a complete rewrite.

```
ELEMENT LENGTH=  10.109589  FEET
ELEMENT DIAMETER  .808767123  INCHES
Y VS X COORDINATES, INCHES
X          DIR        ANT        REF
 0        -13.11       0         12.21
 4.04     -11.59      3.73       22.17
 8.08      -7.82      11.06      31.98
 12.13     -3.32      17.37      36.39
 16.17      .84       21.7       38.44
 20.21     4.25       24.53      39.51
 24.26     6.92       26.41      40.13
 28.3      8.96       27.68      40.52
 32.35     10.52      28.57      40.77
 36.39     11.72      29.22      40.95
 40.43     12.66      29.7       41.08
 44.48     13.41      30.07      41.18
 48.52     14         0          0
 0         0          0          0
 0         0          0          0
 0         0          0          0
 0         0          0          0
 0         0          0          0
 0         0          0          0
 0         0          0          0
```

fig 3A. Tabular output of the computer program for a frequency of 147 MHz. X and Y are normal Cartesian coordinates. The 0's indicate that the end of the element has been passed. These values may be scaled for other frequencies.

The program first determines whether hard copy is needed, then requests its only input, the design frequency. Element length and diameter are then outputted, followed by a table of X, Y values that define the center-line position of each element. The feedpoint, or center of the radiator is taken as the coordinate origin. **Figure 3A** shows the screen presentation (the ending 0's indicate that the end of the element has been passed). Pressing the space bar produces a plot of the lines defining the element centers, as shown in **fig 3B**. Pressing the space bar again either initiates a hard copy or terminates the program.

The general resemblance of this type of antenna to a conventional Yagi is apparent in the figures. The element shaping causes a taper towards the forward direction, even though the elements are the same length. And the deep V of the director gives an effective wide spacing for the director.

high gain is achieved

The performance of this optimized design is very good. According to Chang and Cheng,[1] gain calculates to be 11.8 dBi. Beamwidth is 32 degrees in the element plane, and 62 degrees at right angles to it. Front-to-back ratio is just less than 15 dB in both planes. Feed impedance of the 3/2 wavelength radiator is calculated to be 14 + j33 ohms.

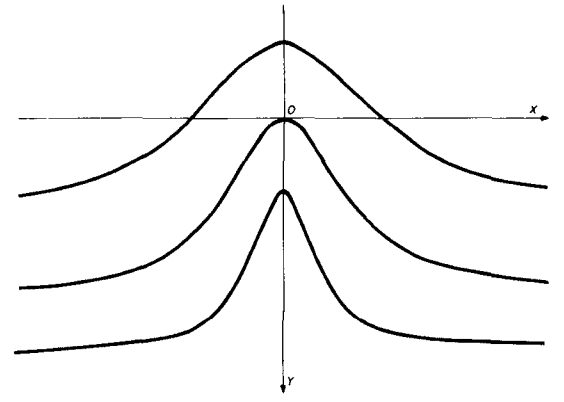


fig. 3B. High resolution computer plot of the data of fig. 3A, showing the center line of the elements. Although each element is 1.5 wavelength long, the shaping makes the reflector appear longer and the director shorter than the radiator.

It should be noted that the design values are optimum only for the element diameter given. This was arbitrarily set at 0.01 wavelength by Chang and Cheng. Performance should not be greatly affected by a reasonable change in element diameter.

Because of the complexity of the required calculations, and the many hours of mainframe computer time necessary to perform them, it is unlikely that there will be much further analysis of the type. Further work will have to be experimental. None has been attempted by the author, but it would seem that additional gain could be secured by placing additional directors of similar shapes in front of the present single director, using appropriate spacings. It would also seem that any of the common matching methods would be usable. Stacking spacing rules of high-gain Yagi type would appear necessary.

conclusion

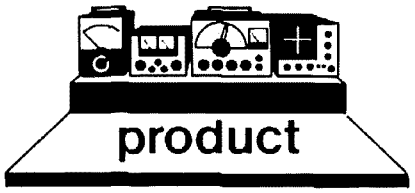
Those who do not have a computer available, or who wish to avoid the tedium of typing in the program, can use these results by simple frequency scaling. All table dimensions should be multiplied by the ratio, 147/new frequency, since the table was calculated for 147 MHz. Element diameter and length vary in the same way.

references

1. Chang-Hong Liang and David K. Cheng, "Directivity Optimization for Yagi-Uda Arrays of Shaped Dipoles," *IEEE Transactions on Antennas and Propagation*, AP-31, Volume 31, No. 3, May, 1983, pages 522-525.

2. F. M. Landstorfer, "A New Type of Directional Antenna," *Antennas and Propagation Society International Symposium Digest*, IEEE, 1976, pages 169-172.

ham radio



REVIEW

RTTY/CW computer interface

Quite a while ago, I quit operating RTTY and retired my old Model 15 because my ham shack wasn't all that soundproof. A few years ago, articles and ads appeared about RTTY computer interfaces. And that meant quiet . . . and no paper! But it wasn't until just a few weeks ago that I finally got an MFJ Model 1224 RTTY/CW Computer Interface Unit. I had a VIC-20, bought for another purpose, and so in less than an hour, I was on RTTY again!

Much of that time was taken finding a power supply to run it. There's a lot of information in the Owner's Manual, but not how much current was needed. A look at the diagram showed only a few chips and transistors. So I took a chance and hooked up a couple of 6-volt lantern batteries and measured it. It turned out to be a little over 200 mA.

specifications

Size: 8.5" W x 6.5" D x 1.5" H.
Power required: 12 volts DC at 250 mA.
Computer: C-64 or VIC-20.
RS-232 interface: available (MFJ-1223).
Software: MFJ-1252 or MFJ-1264 (works with other software).
Accessories supplied: cable with 5-pin plug; 4-pin microphone plug; RCA interface cable.

installation

The unit comes with a cable to connect to the VIC-20 or the C-64. Two clip leads and a Radio Shack cable with RCA phono plugs were used to connect up the audio from the rig. It took a few moments to load the cassette program into the VIC-20 and a few more to get the hang of tuning in the RTTY signal. But right there on the screen was the first RTTY QSO I had "listened" to in years.

operation:

Because the MFJ-1224 doesn't have all the bell and whistles of some of the more expensive units, it takes a little while to get the hang of tuning. Instead of using a 'scope with the familiar "plus" pattern, two LED tuning indicators are used. One is marked "PHASE LOCK" and the other "DATA."

The receiver should be on LSB for receiving RTTY. Most RTTY signals are 60 WPM and 170 Hz shift. Tune from the low tone to a high tone. Watch the phase lock LED. It will flicker and then lock on.

If the station is sending at 60 WPM, the QSO should appear on the screen. If your receiver has an IF shift control, adjust it so the signal is centered in the passband.

switching from RTTY to CW

The RTTY and CW programs are on the same tape. So to switch to CW, you only have to load in the other program, and start tuning. On CW both tuning LEDs flicker with the signal. The MFJ-1224 uses the 2125-Hz RTTY filter in the CW mode. Because 2125 Hz is down the slope of the pass band of most modern-day receivers, IF shift control is almost a necessity for receiving weak CW signals.

transmit

The instructions for using the MFJ-1224 on CW or RTTY transmit are quite complete. Audio Shift is used on RTTY; either Grid Block or Direct Keying can be used for CW.

software

MFJ has software available for both the VIC-20 and the C-64. But they also provide a listing showing the DIP switch positions for most of the other popular RTTY software packages as well. An RS-232 interface (MFJ-1223) is available.

Purchasers of this unit need some experience in RTTY to use it to its full potential, but it is a relatively low-cost way of trying out this mode before leaping in with both feet. The unit *is* well built; I accidentally dropped mine on the basement floor!

For information contact MFJ Enterprises, Box 494, Mississippi State, Mississippi 39762.

—VE3ZL

Circle #302 on Reader Service Card.

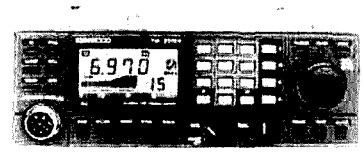


new 2 meter transceivers from Kenwood

Trio-Kenwood Communications has announced the all-new TM-2570A – the first compact 70-watt, 2-meter FM mobile transceiver – and the TM-2550A and TM-2530A 2-meter FM transceivers. The 25-Series includes many new features never before seen in 2-meter FM equipment. All three models have a built-in telephone number memory and automatic dialer. Up to 15 seven-digit telephone numbers may be stored.

All front panel controls, including the 16-key DTMF pad, are back-lighted for nighttime visibility. Twenty-three memory channels store frequency, offset subtone, and telephone number. The CTCSS encoder is programmable from the front panel when the optional TU-7 subtone unit is installed. All standard EIA tones are included, plus the Motorola 97.4 Hz tone, for a total of 38 separate CTCSS tones.

Frequencies are entered into either the VFO or memory with direct keyboard entry. The 25-Series includes frequency coverage for MARS and CAP operation and is modifiable to cover 141-151 MHz. Programmable scanning with priority alert and center stop tuning are standard features.



Digital Channel Link (DCL), a revolutionary new signalling concept compatible with Kenwood's Digital Code Squelch (DCS) system, is available as an option. DCL enables the 25-Series radio to automatically switch to an open channel. Practically speaking, this feature will allow you to automatically QSY to an open simplex channel after making initial contact via repeater.

Supplied accessories include a hand-held microphone with up/down frequency controls, DC cable with fuse, mounting bracket, and microphone hanger. Optional accessories include the PS-50 heavy power duty supply for the TM-2570A, PS-430 DC supply for the TM-2550A and TM-2530A, TU-7 38-tone CTCSS encoder, VS-1 voice synthesizer, MU-1 DCL modem unit, SP-50 deluce mobile speaker, SP-40 compact mobile speaker, CD-10 call sign display, SWT-1 compact antenna tuner, and a wide variety of other station accessories.

For further information, contact Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220.

in-line GaAsFET preamp

Hamtronics, Inc. has announced a new low-cost preamplifier designed to be operated in the antenna line of VHF or UHF transceivers. The new model LNS (low noise switching) preamp is patterned after the popular LNG series, which was the first of the affordable GaAsFET preamps on the market. The heart of the unit is a stable, dual-gate GaAsFET amplifier combined with two special low-loss UHF relays, which use gold-plated contacts for long life. A microstrip PC board combines with these special relays for low VSWR on the transmit throughput.

The preamp is switched out of the signal path automatically whenever a transmit signal is

present. It may also be remotely bypassed manually as desired. The LNS is designed for base or mobile operation, and mounting brackets are provided to allow tower mounting. The LNS can be used with any transceiver up to 25 Watts; and if a separate PA is used, the LNS can be used between the transceiver and the PA. The preamp works with any mode: FM, SSB, CW, ATV, etc. A delay in the RF sensing circuit prevents relay chatter on SSB or CW.

Typical gain of the preamp is 18 dB and typical noise figure is 0.8 dB. Transmit signal attenuation is only 1/2 dB. An LED indicates when the preamp is active. The unit is housed in an attractive aluminum case only 3-7/8"W x 2-7/8"D x 1-1/2"H.

The LNS Transceiver Preamp is available in the following models:

model	tuning range
LNS-144	120-150 MHz
LNS-160	150-180 MHz
LNS-220	200-240 MHz
LNS-432	400-500 MHz

3 dB bandwidth
± 5 MHz
± 10 MHz
± 12 MHz
± 15 MHz

The price of the LNS Transceiver Preamp is only $68 in kit form and $98 wired and tested.

A complete 40-page catalog describing this and other Hamtronics products is available from Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535. (Add $1.00 for first-class mailing; for overseas mailing, please send $2.)

new packet video helps get you started

Would you like to learn more about packet radio? Are you having a hard time getting started up in packet? Well, help is on the way, thanks to Kantronics' first instructional video, *Basic Packet.*

In the course of answering service calls and talking to Amateurs around the country, Kantronics — manufacturers of the new KPC Packet Communicator, an AX.25 Version 2.0 TNC — found that many operators were having difficulty getting started on packet. The difficulties were not with the equipment, but rather with understanding the basic operating procedures of packet radio itself.

Conducted by Phil Anderson, W0XI, the video covers basic subjects such as initial installation and hook-up, VHF and HF operation, digipeating, parameter perming, and some command demonstrations. The tape begins in the classroom, then takes you into the shack for on-the-air demonstrations.

The tape is available to individuals and clubs in both VHS and BETA formats. Suggested retail is $22.50 plus $2.50 shipping and handling. Clubs only can receive a $10 refund if the tape is

Antennas

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416TB	16 el. OSCAR 435 MHz	56.00
A144-10T	10 el. OSCAR 145.9 MHz	49.00
40P-1	OSCAR pack 2mt & 70cm	140.00
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ARX-2	2mt. vert. rngo ranger	30.00
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RM20/RM20S	std. & super resonator	15.95/21.95
RM30	30mt. std. resonator	16.95
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RM75/RM80	75 or 80 std.	18.95
RM75S/RM80S	75 or 80 super	36.95
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SSM-2	stainless ball mt.	17.95
SSM-1	stainless ball & spring mt.	32.95
QD-1	quick disconnect	13.95
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TH3JRS	3 el. 750W pep	218.95
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14AVQ/WBS	4 band trap vert	74.95
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CES RepeaterMaker™

Communications Electronics Specialists, Inc. (CES), has announced the availability of the new CES RepeaterMaker, which allows two mobile radios to be used in a repeater configuration. Features include adjustable hangtime, adjustable time-out timer (TOT), "Roger" or courtesy beep, remote repeat inhibit input and front panel switch, auxiliary PTT relay, inputs for half and full duplex autopatch, connections for CTCSS decoder, LEDs for power, COR,and PTT, all in a compact, attractive case. The CES RepeaterMaker is ideal for use in establishing a primary repeater installation or a back-up repeater utilizing conventional two-way mobile radios.

For more information, contact CES, Inc., 803C S. Orlando Avenue, Winter Park, Florida 32789.

Circle #304 on Reader Service Card.

Yaesu in-line SWR meters

Yaesu Electronics Corporation has released two new in-line SWR and Power Meters. The YS-60 measures both average and peak power output, reflected power, and VSWR in the range from 1.6 to 60 MHz. The YS-500 performs the same measurements covering 140 to 525 MHz range. Three functions provide monitoring of either forward or reflected average transmitter output power for CW, AM, FM, and FSK modes, and VSWR for testing the performance of antennas. The efficient linear circuit design assures accurate measurements with minimum insertion loss over the entire specified frequency range, even at low power levels.

For further information, contact Yaesu Electroncis Corporation, 17210 Edwards Road, Cerritos, California 90701.

Circle #303 on Reader Service Card.

87 STAR for Mac, IBM

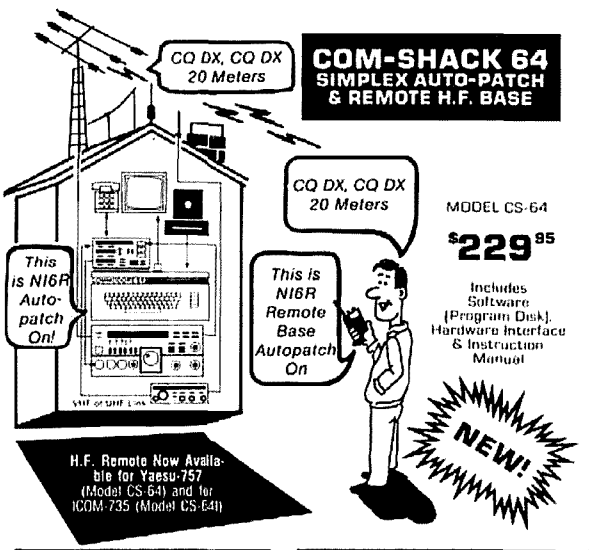
Circuit Busters' popular RF and microwave analysis and optimization program, STAR, is now available for the Apple Macintosh and the IBM PC/XT/AT/Jr with floating point coprocessor. The Macintosh version works with either the 128 or 512K machine and the Imagewriter

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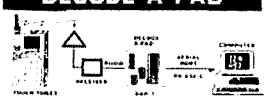


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NEW products

printer. 87STAR for the IBM PC/ST/AT requires the 8087 or 80287 coprocessor, but runs six to seven times faster than the earlier version. 87STAR includes the older version for users who don't have the coprocessor yet installed.

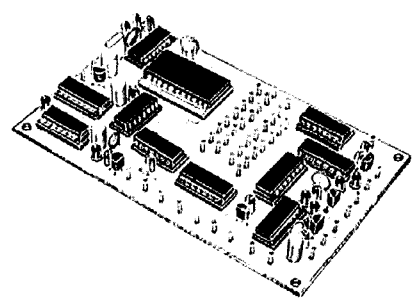
STAR is still available for the Apple II+, IIc and IIe, the Commodore 64, and the Kaypro 2/2X/4/10 CP/M 2.2. 87STAR is priced at $195.00; all other versions are priced under $100.00.

For information, contact Circuit Busters, Inc., 1750 Mountain Glen, Stone Mountain, Georgia 80087

Circle #305 on Reader Service Card.

tone decoder controller module

The new TD-2 DTMF decoder/controller from Hamtronics introduces many new and advanced features. The heart of the TD-2 is a new central-office grade, 16-digit, crystal-controlled DTMF decoder chip with built-in filters to prevent falsing on noise or voice signals. Reliable operation is maintained for any audio input level from 100 mV to 2V p-p. Low power consumption is made possible by use of CMOS circuitry throughout. Power consumption is only 15mA, and the module operates on 10 to 15VDC, with an on-board regulator.



The TD-2 provides five latching on-off functions and up to 12 momentary outputs. This means, for example, that you can control a repeater and autopatch, three other latching functions such as sub-audible tone activation, and several momentary functions. Latching outputs can drive solid-state circuits directly, sinking up to 50mA, or they can drive small relays to switch power loads. The 12 momentary outputs provide a standard 5V logic output.

Codes are easily changed with jumpers; there are no program plugs to replace or eproms to burn. For each function, you have the choice of using a high-security four-digit password or a convenient single-digit command. The TD-2 can be used with any tone source, including radio receivers and telephone lines (with autopatch board). When used with an autopatch, the TD-2 has a built-in "0" and "1" toll call restrictor at no extra charge. A data strobe output, used with the autopatch module, prevents retransmission of confidential control tones.

The TD-2 is suitable for remote control of repeaters, autopatches, and subaudible tone decoders; for selective calling, industrial control, telemetry, computer interface, and other applications. LEDs indicate latch circuits automatically reset to default values on power up. The size of the double-sided PC board is only 3 x 5 inches, so it fits in easily with other equipment.

The TD-2 sells for $110 in kit form and $160 wired and tested. Full documentation is provided with helpful application data.

A complete 40-page catalog of Hamtronics products is available for $1.00 from Hamtronics, Inc., 65-F Moul Road; Hilton, New York 14468-9535. (For overseas mailing, please send $2.00)

new signalling concept

A revolutionary new selective calling/selective linking system has been developed by Kenwood for Amateur Radio use. Called Digital Channel Link, or DCL for short, its many features include automatic connection, frequency recall, vacant channel location, and selective calling of individual transceivers or groups of transceivers.

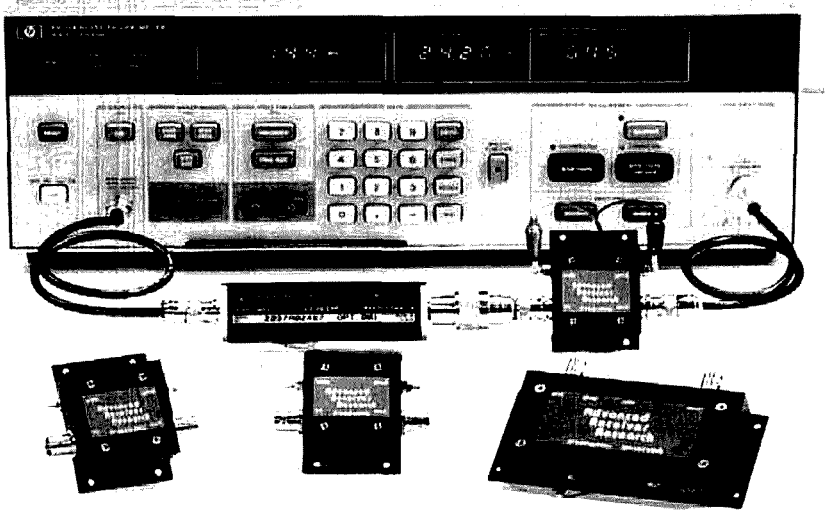
Here's how it works: the DCL system searches for an open channel, remembers it, returns to the the original frequency and transmits control information to the DCL-equipped station that switches both transceivers to the open channel. High speed digital information allows the whole process to take only an instant.

In addition to this selective linking feature, DCL can also be used for selective calling, similar to Kenwood's DCS system: a five-digit code group is sent which opens squelch on a DCS transceiver with the matching code. Additionally, a six-character burst of ASCII is sent. Station call signs are normally programmed into this ASCII portion. The CD-10 Call Sign Display unit, which can be used with any receiver, may be used to display the transmitted DCS or DCL (ASCII) call sign. The CD-10 can store up to twenty incoming call signs for monitoring and logging purposes.

The digital Channel Link system should add more convenience to repeater operations. Using DCL, it becomes a simple matter to QSY to an open simplex channel after making initial contact via repeater.

For details, contact Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220.

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P50VD	50-54	<1.3	15	0	DGFET	$29.95
P50VDG	50-54	<0.5	24	+12	GaAsFET	$79.95
P144VD	144-148	<1.5	15	0	DGFET	$29.95
P144VDA	144-148	<1.0	15	0	DGFET	$37.95
P144VDG	144-148	<0.5	24	+12	GaAsFET	$79.95
P220VD	220-225	<1.8	15	0	DGFET	$29.95
P220VDA	220-225	<1.2	15	0	DGFET	$37.95
P220VDG	220-225	<0.5	20	+12	GaAsFET	$79.95
P432VD	420-450	<1.8	15	−20	Bipolar	$32.95
P432VDA	420-450	<1.1	17	−20	Bipolar	$49.95
P432VDG	420-450	<0.5	16	+12	GaAsFET	$79.95
Inline (rf switched)						
SP28VD	28-30	<1.2	15	0	DGFET	$59.95
SP50VD	50-54	<1.4	15	0	DGFET	$59.95
SP50VDG	50-54	<0.55	24	+12	GaAsFET	$109.95
SP144VD	144-148	<1.6	15	0	DGFET	$59.95
SP144VDA	144-148	<1.1	15	0	DGFET	$87.95
SP144VDG	144-148	<0.55	24	+12	GaAsFET	$109.95
SP220VD	220-225	<1.9	15	0	DGFET	$59.95
SP220VDA	220-225	<1.3	15	0	DGFET	$67.95
SP220VDG	220-225	<0.55	20	+12	GaAsFET	$109.95
SP432VD	420-450	<1.9	15	−20	Bipolar	$62.95
SP432VDA	420-450	<1.2	17	−20	Bipolar	$79.95
SP432VDG	420-450	<0.55	16	+12	GaAsFET	$109.95

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DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

SCHEMATICS: Radio receivers 1920s/60's. Send brand name, model No., SASE. Scaramella, PO Box 1, Woonsocket, RI 02895-0001.

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BACK ISSUES HR Oct 71 thru Dec 81. 72-77 in binders. $65. VE3BJY, B. Brenko, 2 Vista Dr., Foothill, Ontario L0S 1E0

DIGITAL AUTOMATIC DISPLAYS. All transceivers. Six 1/2" digits. 5" wide x 1-1/4" cabinet! Send $1.00 for information. Receive a $25.00 discount. Grand Systems, POB 2171, Blaine, Washington 98230.

WANTED: Service and operating manual for Gonset GSB-100 model 3233 transmitter. Also wanted a Viking Challenger transmitter that is complete. Fadden, Box 1223, Blaine, WA 908230.

THE GOOD SAM HAMS invite RV operators to check in the Good Sam Ham net 14.240 *Sundays 1900Z also 3.880 Tuesdays at 2359Z. Net control N5BDN, Clarksville, IN*

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OLD RADIO transcription discs wanted. Any size, speed. W7FIZ, Box 724 HR, Redmond, WA 98073-0724.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $10.00. Beginners RTTY Handbook $8.00. PO Box RY, Cardiff, CA 92007.

UHF PARTS. We stock GaAs Fets, Trimmers, and many of those impossible to find parts for the UHF builder. For example: MGF 1202 at $10.00, Finger Stock at $7.50 per 16" strip. SASE brings list, Microwave Components, 11216 Cape Cod, Taylor, MI 48180.

IMRA, International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, New York 10538.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

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CUSTOM MADE EMBROIDERED PATCHES. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept 301, 4202 N. Drake, Chicago, IL 60618.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

PRINTED CIRCUIT BOARDS and kits for QST articles. Call or write for information. A&A Engineering, 7970 Orchid Drive, Buena Park, CA 90620. (714) 521-4160.

PACKET/ASCII/BAUDOT/CW for IBM-PC. SASE to: Emile Alline, 773 Rosa, Metairie, LA 70005.

WANTED NC400 receiver. Any condition. W2PUA, 112 Tilford Road, Somerdale, NJ 08083. (609) 783-4175

CABLE TV CONVERTERS/DESCRAMBLERS. Guaranteed lowest prices in US. Jerrold, Hamlin, Zenith — Many others. Lowest dealer prices! Orders shipped within 24 hours! Complete illustrated catalog, $2.00. Mastercard, VISA, COD accepted. Pacific Cable Co., Inc., 7325-1/2 Reseda Blvd,#1011, Reseda, CA 91335. (818) 716-5914.

DISCOUNT CABLE TELEVISION EQUIPMENT. We have IN STOCK complete systems for Hamlin, Oak, Jerrold and other cable systems. Please send for a free information package — it could save you big money on your purchase of these and other CATV items. Easy View, PO Box 221, Arlington Heights, IL 60006 (312) 952-8504. Ask for Rudy Valentine.

HAPN (Hamilton and Area Packet Network) announces the availability of a packet radio terminal node controller (TNC) on a card that plugs into a slot in an IBM-PC or compatible computer. Card contains terminal node controller and 1200 baud 202-type modem. Card available as bare board ($75 US) or assembled and tested unit ($199 US). AX.25 software included with each order. For order outside N.A. please add $3.00. For information/orders: HAPN, Box4466, Station D, Hamilton, Ontario, Canada L8V 4S7.

CHASSIS and cabinet kits. SASE K3IWK, 5120 Harmony Grove Road, Dover, PA 17315.

ELECTRON TUBES Radio, TV & Industrial Types — Huge Inventory. Send for 80% off tube listing. Call Toll Free (800) 221-5802 or write Box HR, Transleteronic, Inc., 1365 39th Street, Brooklyn, NY 11218 (718) 633-2800.

WORLD TOP BAND Frequency Allocations Listing. Handy "Top Band" operating and contains all currently available CW and SSB allocations for countries on ARRL DXCC list. (317 total, 10 no authorization, 61 no information available). Also includes OST sent/received column for keeping track of cards. $5.50 USA, Canada and Mexico. $7.50 elsewhere. Dennis Peterson, N7CKD, 4248 A Street SE, Space 609, Auburn, WA 98002.

WANTED: Heath Curve Tracer IT-3121 kit/assembled. Mission School needs 10 units. Call Manila, P.I. 832-1773 or write Mission POB 1651 MCC Makati MM, P.I. SALE: HW101 with power supply - manual. Best offer over 150. Single bander (20m) offer. Both prof. wired. Excellent condition. W2PUA, 112 Tilford Road, Somerdale, NJ 08083. (609) 783-4175.

CAPS with call/handle. Mesh or foam backs. 12 colors/white fronts; brown/tan. Mens/ladies styles. $7.50 net. CERAMIFUN 315 — 11th Street, Ramona, CA 92065 (619) 789-4730. Visa/MC accepted. Quantity rates to clubs.

FOR SALE: Knight 50W transmitter. Matching VFO. 10-80 meters. $75.00. Collins Type 32V Transmitter $250.00. Donald M. Cox, 318 Park Blvd., Orange Cove, CA 93646. (209) 626-4219.

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CLEANING out Ham Shack. 35 years of receivers, transmitters, accessories, tubes and parts. Send SASE for 6 page list. W9VZR, 4627 North Bartlett Avenue, Milwaukee, WI 53211.

REPEATER FOR SALE: Spectrum model SCR-1000 2 meter, 2-1/2 years old, freshly factory reconditioned and upgraded. John L. Hackman, WB4VVA, PO Box 194, Mount Pleasant, MI 48858.

QST July 1985 issue wanted. Prof. G. Felser, Trazerberggasse 5, 1130 Vienna, Austria.

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2 METER AMP KITS: 8877 legal limit kit $395. 3CX800A7 900W kit $325. Also HV power supplies, CX600N relays, parts and EME newsletter. SASE for catalog. 2 Meter EME Bulletin, 417 Staudaher St., Bozeman, MT 59715.

CABLE TV CONVERTERS & EQUIPMENT: Plans and parts. Build or buy. SASE for information. C & D Electronics, PO Box 1402, Dept. HR, Hope, AR 71801.

RIW-19 432 MHz beams by K3IPW. *SASE for information.*

COMING EVENTS

Activities — "Places to go . . ."

CALIFORNIA: Flea Market/Boneyard Sale. Foothill College, Los Altos Hills. March-Sept. 2nd Saturday of every month. 7 AM Sellers. 7:15 AM buyers. Talk in 145.27(-) or 147.570 simplex. FCC exams (408) 255-9000.

GEORGIA: The 7th annual Lake Hartwell Hamfest, sponsored by the Anderson, Hartwell and Toccoa Amateur Radio Clubs. May 17 and 18, Lake Hartwell Group Camp, Hwy 29, Hartwell. *Free admission, free camping and free flea market space.* Activities for the entire family. For information: Merrick A. Counsell, W2BNS, 215 Nottingham Way, Anderson, SC 29621.

ILLINOIS: The Starved Rock Radio Club Hamfest, June 1, Princeton. Same place as last year. SASE please for complete registration materials, map, etc. SRRC, W9MKS, RFD 1, Box 171, Oglesby, IL 61348 (815) 667-4614

OHIO: The Medina County Hamfest, May 11, Medina County Community Center Building, 735 Lafayette Rd., Medina. Sponsored by the Medina 2 Meter Group. 8 AM to 2 PM. Building and flea market setup 6 AM. Tickets $3.50 advance, $4.00 at the door. Tables $6.00. Flea market space $4.00. Talk in on 147.63/03, K8TV/R. For tickets, table reservations: PO Box 452, Medina, Ohio 44258. (216) 725-4492 or (216) 769-3033.

ILLINOIS: The Moultrie Amateur Radio Klub Hamfest, April 20, Coles County Airport, Mattoon. Contact MARK, PO Box 79, Sullivan, IL 61951.

CALIFORNIA: FCC exams, Novice-Extra. Sunnyvale VEC ARC. (408) 255-9000 24 hour. 73, Gordon, W6NLG, VEC

OHIO: The all new 17th Annual B*A*S*H — New Location, new entertainment, new food — will be held *on Friday night* of the Hamvention, April 25, 1986. The new location is in the Conference Center at the HARA Arena and Exhibition Center, (the same location as the Hamvention) starting at 7 PM. There is no admission charge, and free continuous entertainment. Food and beverages available. Two exciting top awards, and many others. Stay right at HARA when the Hamvention closes on Friday evening and meet your friends and join us for an evening of fun and entertainment. Sponsored by the Miami Valley FM Association, POBox 263, Dayton, Ohio 45401.

OKLAHOMA: The Great Plains ARC's 5th annual Northwest Oklahoma Eyeball & Swapmeet. Sunday, April 13, 9 AM, *Mooreland.* Admission *$2.00 at door. Dealer and swap tables* available no charge. VE tests Saturday, April 12. Campsites available. Talk in on 147.72/12, 146.13/73 and 146.52 simplex. For information: (405) 994-5394 or 994-5453. Write: NR5L, Gordon Richmond, Rt. 1, Box 12, Mooreland, OK 73852 or N5CCV, Gerald Bowman, Box 356, Mooreland, OK 73852.

COLORADO: The Aurora Repeater Association will hold its annual ARA Swapfest, Sunday, April 27, National Guard Armory, 55 S. Potomac, Aurora, 8 AM to 3 PM. Tables, raffle, FCC exams. For information: Aurora Repeater Assoc., PO Box 31043, Aurora, CO 80041 or call John(303) 344-1915.

INDIANA: *The third annual Columbus Amateur Radio Club* Swapfest, Saturday, April 5, 9 AM to 5 PM, 4H Fairgrounds, SR 11 South, Columbus. Talk in 146.79 repeater. For reservations: Chuck Roberts, 2950 S. Lake Drive, Columbus, IN 47203.

NEW JERSEY: TCRA Hamfest Tri-County Radio Association, Sunday, May 4, Passaic Valley Community Center, off Valley Road, Stirling. 9 AM to 3 PM. Tables $7. AC $10. Registration $2. Limited tailgating by reservation only. All reservations W2EUF, Dick Franklin, Box 182, Westfield, NJ 07090. (201) 232-5955.

OHIO: The Portage Amateur Radio Club's Hamfair, Sunday May 18, Randolph Fairgrounds, Ravenna. Gates open 6 AM for dealers, 7:30 for public. Indoor/outdoor flea market. ARES, ARRL, DX, packet and computer forums. Tickets $3.00 advance, $3.50 at the gate. Mobile check in 144.79/145.39 repeater. For tickets send check and SASE to PARC, c/o Joanne Solak, KJ3O/8, 9971 Diagonal Road, Mantua, Ohio 44255. For information call (216)274-8240.

LOUISIANA: BRARC Hamfest, May 3 and 4, Baton Rouge. Free admission. VE exams Saturday and Sunday, 8:30 AM. 30 day advance registration. Send SASE, Form 610 and check for $4.25 payable to ARRL/VEC to George Perry, W5LVX, 17424 Lady Constance, Greenwell Springs, LA 70739. Some walk-ins. For further info SASE to Rick Pourciau, NV5A, 879 Castle Kirk, Baton Rouge, LA 70808.

ARKANSAS: The Northwest Arkansas ARC will hold its 6th annual Hamfest, Saturday, May 3, Rogers Youth Center, 315 W. Olive Street, Rogers, AR. 8 AM to 4 PM. Exhibitors and flea market tables $2.00 per space. Doors open 6 AM for exhibitors only. General admission free. Talk in on 16/76, 63/03 and 52 simplex. For more information: Roy Milliren, AF5W, 2014 S. 16th Street, Rogers, AR 72756.

CALIFORNIA: The Fresno Amateur Radio Club will hold its 44th annual Hamfest, May 2, 3 and 4. Airport Holiday Inn. FCC exams will be given. DX and Emergency programs. Forums and demonstrations. Talk inon 146-34-94. For information: Harry H. Billings, Fresno ARC, PO Box 783, Fresno, CA 93712. (209) 268-6314.

INDIANA: The Putnam County Amateur Radio Club's 4th annual Hamfest and Auction, April 12, Putnam County Fairgrounds, north of Greencastle. Setup time 6:30 AM. Doors open to public 8:00 AM. Auction 1:00 PM. Admission $3.00. Children under 12 free. Flea market tables $3.00 each. Your tables $2.00 each. Talk in 147.33/93. For information or table reservations SASE to Kent Douglas, K9JCR, RR4 Box 586, Greencastle, IN 46135 (317) 672-8237 or Nick Aubrey, N9FCB, RR2, Box 592, Greencastle, IN 46135 (317) 653-5290.

MISSOURI: The PHD Amateur Radio Assn.'s annual State ARRL Convention, April 11-13, old Kansas City Airport, north of downtown KC. Registration $4.00 (good for all 3 days). Swap tables $10.00, includes one registration. Saturday night banquet $10.50 by advance registration. VE exams. Talk in on 34/94 repeater. For information/registration: PHD Amateur Radio Assn, Inc., PO Box 11, Liberty, MO 64068-0011 (816) 781-7313 or 452-9321.

WISCONSIN: The Ozaukee Radio Club's 8th annual Cedarburg Swapfest, Saturday, May 3, 8 AM to 1 PM, Circle B Recreation Center, Highway 60 and County I, Cedarburg. Admission $2.00 advance, $3.00 at the door. 4' tables $3.00 each. Sellers setup 7 AM. For tickets, table reservations or more info SASE to ORC Swapfest, 101 E. Clay St., Saukville, WI 53080.

NEW YORK: The Suffolk County Radio Club's Indoor/Outdoor Flea Market, Sunday, May 4, 8 AM to 3 PM, Republic Lodge No. 1987, 585 Broadhollow Road, Melville, LI. General admission $2.00. Wives and children under 12 free. Indoor tables $7.00. Outdoor space $5.00. Each includes one admission. Talke in on 144.a61/145.21 and 146.52. For more information: Bill Sullivan, N2ETG (516) 689-9871 evenings.

NEW YORK: Manhattan's Quarterly Computer Show and Electronic Flea Market, April 5, 10 AM to 4 PM, Christ Church Auditorium, Park Avenue and East 60th Street. And, again on April 26, 10 AM to 4 PM, St. Anthony's Church Community Hall, Sullivan and Houston Streets. Admission to each show $2.50. For more information: Mr. Johnson. Public Domain Software Copying Company, 33 Gold Street, New York, NY 10038. (212) 732-2565.

NORTH CAROLINA: Raleigh, the City of Oaks and the Raleigh Amateur Radio Society presents the 14th annual RARS Hamfest, NC State ARRL Convention and Computer Fair, Jim Graham Building, NC State Fairgrounds, Hillsborough Street. Advance registration $3.50 until April 7. $5.00 at the door. Flea market space, one table and 2 chairs (ours only) $6.00 each. FCC exams by pre-registration prior to April 1. Contact John Johnson, WM4P direct. Free welcoming party in Graham Building Saturday night, 7 - 10 PM. Wouff Hong Ceremony. Talk in 04/64 and 28/88. For more information: Rollin Ransom, NF4P, 2447 Fairway Drive, Raleigh, NC 27603. (919) 779-5021.

MASSACHUSETTS: The 12th annual ECARA Flea Market, April 27, 10 AM to 2 PM, 200 Sportsmen's Club, Sutton Road, Webster. Tables $5.00 advance; $7.00 at the door. Admission $2.00. Free parking. For information: Tom Francis, KB1SP (617) 743-7283. Dick Spahl, K1SYI (617) 943-4420. Don Amirault, K1APE (203) 923-2727.

MASSACHUSETTS: The Montachusett Amateur Radio Association will hold an indoor flea market, Saturday, April 26, Knights of Columbus Hall, Electric Avenue, Fitchburg. 9 AM to 3 PM. Admission $1.00. Reserved tables $8.00 advance. $10.00 at the door. Refreshments and free parking. Doors open 8 AM for sellers. Talk in on 144.85/5.45 and 52 simplex. For table reservations send check payable to MARA, c/o Jim Beauregard, KB1AY, 7 Mountain Avenue, Fitchburg, MA 01420.

MASSACHUSETTS: Tailgate High Tech, computer and Amateur Radio Flea Market, Sunday, April 20, 10 AM to 4 PM, Albany and Main Street, Cambridge. Admission $1.50. Sellers $5 per space includes 1 admission. Setup 9 AM. Talk in 146.52 and 449.2/444.2, W1XM/R. For space reservations or further information: Jamie (617) 262-5090 or 253-2060. Sponsored by the MIT Electronics Research Society and W1XM/R.

PENNSYLVANIA: The 4th annual Southern Aleghenies Hamfest, Sunday, April 13, 7 AM to 4 PM, Bedford County Fairgrounds, Bedford. Sponsored by the Horseshoe Radio Club, Blue Knob Repeater Assn, Bedford Co. ARC, Mountain ARC and Somerset Co. ARC. Admission $3.00. Tailgating available. Reserved inside tables. Refreshments and a consignment auction. Talk in on 145.49/89, 444.2/9.2 and 146.52 simplex. For information: Gay Rembold, W3DFW, 949 Winifed Rd, Cumberland, MD 21502 (301) 724-0674 or (814) 445-7486.

NEW HAMPSHIRE: Deerfield. The Hosstraders will present their Spring Tailgate Swapfest, Saturday, May 10 at Deerfield, NH Fairgrounds. Admission $2 per person; no extra charge for sellers. Friday night camping at nominal fee, but no entry before 4 PM Friday. Profits benefit Shriners' Hospital. Last year's gift: $11,754.46. Talk-in 146.40-147.00. New England's biggest Hamfest is sponsored by Bob, W1GWU, Joe. K1RQG, and Norm, WA1IVB. For map SASE to WA1IVB, RFD Box 57, West Baldwin, ME 04091.

MASSACHUSETTS: The MIT UHF Repeater Association and the MIt Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday, April 23, 1986,7 PM, MIT Room 1-134, 77Mass Ave, Cambridge, MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-5820/646-1641 or Craige Rodgers at 225-6616. Exam fee $4.00. Bring copy of current license, 2 forms of picture ID and completed form 610 (available from FCC in Boston. 223-6609).

OPERATING EVENTS

"Things to do . . ."

ARMED FORCES DAY: In recognition of the 37th anniversary of this event, Amateur Radio Station W4ODR, located Northside aboard Naval Air Station Memphis, Millington, Tennessee, will be operated by sailors and Marines on Saturday, 17 May from 1400Z to 220Z. For information on W4ODR, NAS Memphis contact: Station Custodian, Chief Petty Officer Bob Donan, KA4FAL, (901) 872-2007.

QRP ARCI Spring CW Contest, April 19 to April 20. For information Eugene Smith, KA5NLY, Chairman, POBox 55010, Little Rock, AR 72225.

ATLANTA, GA: The Metro Atlanta Telephone Pioneers ARC will operate W4OTA, April 18-20 to help celebrate A Taste of Atlanta 1986. Telephone Pioneers nationwide are encouraged to participate. For Special Event QSL SASE with log info to MATPARC/Taste of Atlanta, John C.Parker, PO Box 54017, Atlanta, GA 30308.

Groton, CT: The Radio Amateur Society of Norwich (RASON) will operate Croaker Memorial Special Event Station KA1IFG from 1700ZApril 19 to 1700Z April 20 from the submarine USS Croaker to commemorate the 42nd anniversary of its commissioning. Frequencies: SSB – 3.890, 21.290. CW – 3.730, 7.130, 21.130. QSL with SASE to RASON, PO Box 903,Norwich, CT 06360.

SCHOLARSHIP AWARD: The Atlanta Radio Club is pleased to announce its 1986 scholarship awards program. Two sums of $1250.00 each will be awarded to the winners. Applicants must be licensed Radio Amateurs graduating from high school and entering an accredited college or university as Freshmen for the first time in 1986. Judging is based on school grades, citizenship, ham radio achievements and financial need. For application blanks write: Phil Latta, W4GTS, 259 Weatherstone Parkway, Marietta, GA 30067.

THE FOUNDATION FOR AMATEUR RADIO, INC., a nonprofit organization with headquarters in Washington, DC, plans to award 21 scholarships for academic year 1986-87. Licensed Radio Amateurs may apply for these awards if they plan to pursue a full-time course of studies at an accredited university, college or technical school. For additional information and application form send letter prior to May 31, 1986 to FAR Scholarships, 6903 Rhode Island Avenue, College Park, MD 20740.

Derby and District Amateur Radio, incorporating Derby Wireless Club 1911, will be celebrating its 75th anniversary during 1986. The Society plans at least one event per month throughout the year each from a different location with the City of Derby. The callsign to listen for will be GB3ERD.

1986 marks the 50th anniversary of the Greater Cincinnati Amateur Radio Association. A number of special events are planned. Watch for announcements here.

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THE GUERRI REPORT

by Ernie Guerri, W6MGI

analog vs. digital — the difference blurs

It wasn't too long ago that the line between analog and digital operators (devices that perform functions in their respective domains) was easily distinguished. We accept the idea that analog functions have an "unlimited" number of steps and that digital functions are easily distinguished by the fact that the function takes place in discrete steps. But what do we really mean by "unlimited" or "discrete?" Well, it turns out that for most analog functions, the "limit" is set by the resolution of the relevant human sense — touch, smell, hearing, etc. But in reality we find that the presumption of a continuous gradient of sensory perception has its limits. The average eye can distinguish between 55 and 75 levels of gray. This means that 6-bit digitizers (64 shades) are all that is necessary to present an image that has appears to have continuous tones. This is well within the capabilities of today's digital techniques, and literally thousands of shades and colors are possible for even the most demanding visual applications such as computer-aided design and engineering.

The same is true for the acoustic domain. Much professional recording is now done with equipment that converts the analog signal to a purely digital form. Because the digital signal has better "resolution" than the human ear, it's a simple matter to process the signal for noise reduction, echo, timbre, and other characteristics that would be nearly impossible to correct in analog form. The result is the splendid performance of compact disks and the spectacular recordings possible with the 16-bit pulse code modulation (PCM) adapters that are available for use with VCRs.

Similar strides are being made in the development of tactile and olfactory sensors and processors. The principal applications for these devices are as enhancements for robots on the factory floor. Digital signal processors, combined with motion controllers that "naturally" respond to digital signals (rather than analog signals) are soon to perform the most delicate human tasks, with significant improvement in repeatability. Reliability and availability are issues yet to be resolved. It may well be that as the level of robot complexity increases, the robots may also get "sick," but in ways we have yet to determine.

A practical application for the newest breed of digital signal processors will be the re-emergence of the picturephone. When the first Bell picturephone was implemented over 10 years ago, it offered only black and white images and was limited by a rather poor ability to transmit motion. (The bandwidth compression technique used at that time was called *conditional replenishment* — which meant that only the portions of the picture that had changed were transmitted.) Processing speed limitations kept the image from having the quality that consumers demanded, and, of course, color images were not available.

The same basic coding scheme is now being used by companies in both the United States and Japan to offer video teleconferencing in full-color motion over standard 56-bit data lines. High-speed digital signal processing chips, combined with some enhancements to the original replenishment scheme, have made the desired results possible at costs that are quite affordable for business applications.

microscope can "see" atoms

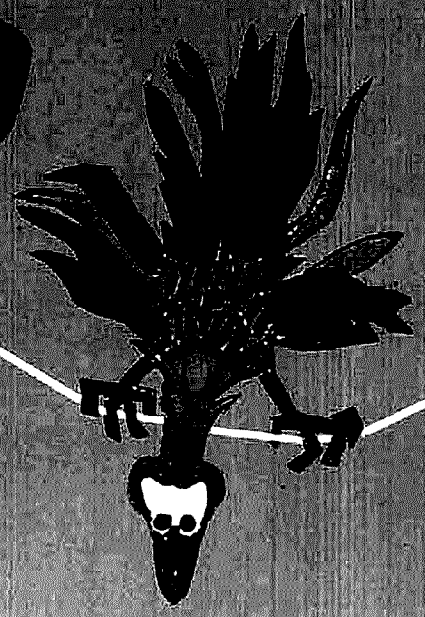
Researchers at IBM Laboratories in Switzerland are refining a recently developed technique that permits a scanning microscope to display the individual atoms of a surface structure. The device operates by scanning a very small probe a few atomic diameters above the surface. An electric field proportional to the distance between the probe and the surface is measured and converted to a visual "image" of the surface. Since the field is proportional to the square of the distance, surface characteristics are easily distinguished from background or "noise." The technique can be used in conjunction with other microscopy methods to serve as a magnifier or zoom device. Typical operation requires a few minutes to scan the surface under examination and produce a three-dimensional image of the atomic structure. Because most surfaces are not uniform for a distance of more than a few dozen atoms, the actual area scanned is quite small. This is not a real limitation, however, since the purpose of the device is extreme magnification of very specific features. The device will have special benefit for those who are working at the limits of semiconductor fabrication technology and are seeking to develop the smallest possible features or achieve the highest levels of integration.

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circulation

cover art:
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REFLECTIONS

communication for fun and profit

Wouldn't you think that after working an eight-hour day you'd do *anything* except continue the same thing you'd been doing all day?

Some people work at jobs that they dislike. Perhaps they're stuck — but for one reason or another, they just can't leave their positions. They wouldn't be caught dead spending even a minute thinking about their work once they're finished for the day. Others enjoy their jobs but make it a point to sharply differentiate between their work and their private lives. Still others don't see this separation so clearly; they either take work home or participate in some aspect of it after hours.

Psychiatrists would probably agree that the middle group is the healthiest and well-balanced. They could earn a living treating the first and third groups.

As editor of a ham magazine I find that I gravitate towards the third group. I spend my daylight hours pushing papers from one side of the desk to the other; miraculously, manuscripts are edited, schedules are developed, correspondence is answered, and magazines get published. But when it's time to quit, I still find myself thinking about a preamplifier that exhibits extremely high IM performance, a multi-element antenna design that can develop a high forward gain and front-to-back ratio, or a multi-stage bandpass filter that can be designed and optimized through the use of a CAD program.

Often I just go down to the radio room to briefly scan the band to see if Europe is coming through on a more southerly route — didn't WWV just announce a *K* index of 5? But wait a second . . . there's LZ1KDP on 3792 describing his three-element wire beam at 200 feet above the streets of Sofia. Trying to picture that scene, I find that I have the microphone in hand and I'm dropping my call several times. Suddenly there's Stefan's voice coming out of the loudspeaker, warmly saying hello from Bulgaria, and we basically continue where we last left off from the previous contact. Pop, (Popa, YU7PFR), comes on frequency and we find ourselves discussing the minor pattern distortion achieved by adding segments at one end only that change the resonant frequency of his director, reflector in his three-element wire Yagi. Perhaps we're discussing Serbo-Croatian music or — back to the technical realm — the merits of the Plessey SL1640 IC in a circuit he's building. In short, we find ourselves going beyond what some perceive as the humdrum nature of today's all-too-common Hello/Goodbye QSO: "Hello, your signal is . . . my QTH is . . . my name is . . . please QSL . . .Goodby."

Just think of the opportunity presented to us. We can communicate with others over thousands of miles, discussing — within the limits of reason and common sense — a myriad of subjects. We can increase our friendships, learn languages, discuss mutual problems we have with a particular computer program. It's almost like having good company over after dinner.

This form of communications is not just fun — I *profit* from this. We all do. For when you extend friendship, hear that laughter from someone who's been trying very hard to say something in English while you try to respond in his language, even though you know only a few words, we all profit. We profit when suddenly we start to observe the makings of a bent propagation path from both ends of the circuit . . . see it develop, peak, and disappear, and see the potential for other exciting contacts as well. We profit by learning to live with each other under crowded conditions, not entirely unlike those of an inner-city tenement, on this cluster of frequencies we call an Amateur band. It's all here to do with as we please. I choose to communicate for fun and profit.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

THE CONCEPT THAT CITIZENS HAVE NO RIGHT OF ACCESS TO THE RADIO SPECTRUM is still very much alive, according to comments by California Congressman Carlos Moorehead at the fourth and final House subcommittee hearing on the Communications Privacy Act of 1985 March 5. Early in the hearing Deputy Assistant Attorney General James Knapp of the Justice Department Criminal Division testified that his agency didn't believe scanning for "recreational purposes" should incur "criminal or civil liability." Instead, the Justice Department position was that it would be a crime only if "the citizen both intercepts and divulges communications under circumstances in which the interception and divulgence are illegal, tortious, or for criminal gain" -- which is pretty much the present law!

However, Congressman Moorehead Severely Criticized The Justice Position, stating "It's very clear there are all kinds of mischievous things you can do if you have one of these scanners" such as hearing "family fights, conversations between someone and their girl friend, inside information that stockbrokers might be giving out over the phone..." He then read from a scanner ad a long list of the things that can be heard -- including Amateur Radio -- noting that "these are private calls and obviously something has got to be done to limit the range of these scanners." He further stated the purpose of the bill was to prevent people "from deliberately trying to intercept any or all of these calls..." though admitting that accidental interception such as through a TV set shouldn't be prosecuted if the listener quickly "switched to the next band."

Following Moorehead's Condemnation Of The Justice Position, Knapp agreed Justice would consider whether "We should go that second step and predicate a violation based upon the interception of 'the radio signal' itself."

"Shockingly Pervasive Misrepresentation" By The Cellular Industry in its previous testimony about cellular privacy before the subcommittee was charged by the Association of North American Radio Clubs (ANARC) in its comments. ANARC emphasized that cellular transmissions are even less private than a cordless phone's, due to the far greater range of cellular transmitters and the variety of means by which cellular can be intercepted.

Subcommittee Work On HR 3378 Is Expected To Be Completed By Late March, at which time it will be taken up by the full Committee and could quickly move to the house floor. At the same time the Senate version of HR 3378 will become active there as well. As the report above demonstrates, this bill's threat to a citizen's traditional access to the radio spectrum as described in Ham Radio's February editorial is still a very real one!

FCC ACTION ON ARRL'S PROPOSAL TO ENHANCE NOVICE PRIVILEGES may very well come in time for the Dayton Hamvention in the form of a Notice of Proposed Rule Making. Most of the comments filed in response to the proposal (RM 5Ø38) were favorable, though some did in fact raise questions about just what it is expected to accomplish, since it would require raising Novice qualifications to correspond with the new privileges. Another concern was that it might decrease incentive to either learn the code properly or even upgrade, since the "enhanced" Novice license would provide access to both HF and VHF phone bands.

THE COUNCIL FOR AMATEUR RADIO EXAMINING IS NOW OFFICIALLY recognized as a "Not For Profit" corporation under Illinois law, and efforts are under way to establish its tax-exempt status for donations as well. Representatives of the CARE member VECs, which now include all national VECs (except the ARRL) as well as many regional VECs, plan to hold an informal strategy meeting during the Dayton Hamvention weekend.

"WESTERN INTERNATIONAL COORDINATOR'S ASSOCIATION" IS THE NEW NAME tentatively chosen for the "Pacific Area Coordination Association" announced in last month's Presstop. The new name was adopted to reflect the group's expanded sphere of influence, now planned to extend to Alaska and the western portions of both Canada and Mexico. A formation meeting has now been set for September 6, during the ARRL National Convention in San Diego.

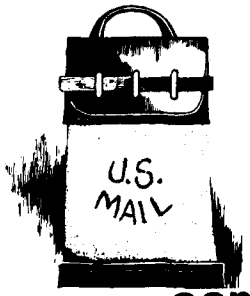
APPARENT ON-THE-AIR "BUSINESS" ACTIVITES ON "FLORIDA TRADERS" 75-meter swap net led to issuance of FCC Notices Of Violation to 22 Amateurs in the Southeast. The notices were sent out by the Ft.Lauderdale Field Office after it monitored two net sessions, and went only to Amateurs who quoted prices on the air or whose "stock" gave the appearance that they were running a business rather than disposing of excess personal equipment.

Though Swap Nets Per Se Are Not Prohibited By The Rules, "business" activities on the Amateur bands are. This crackdown, which came about as a result of numerous complaints from Amateurs about this particular net's activities, does not herald any FCC change in direction or indicate the beginning of a nationwide FCC effort.

THE FCC RELAXED ITS PROHIBITION AGAINST "AUTOMATIC CONTROL" for third party traffic relays in a waiver issued March 14. The third party traffic problem had arisen in the FCC's Report and Order on PR Docket 85-1Ø5, which permitted automatic control of digital communications above 5Ø MHz but pointed out that the rules still required a control operator's presence when third party traffic was being handled -- and that any traffic not originated by the relaying station was considered "third party."

The Waiver Applies To Packet Communications Using AX.25 Protocol, and remains in effect until the Commission can consider a number of Petitions For Reconsideration that were filed in PR Docket 85-1Ø5. It was granted in response to a Petition for Extraordinary Relief filed by the ARRL. FCC review is likely late this year.

THE 12TH ANNUAL EASTERN VHF/UHF CONFERENCE WILL BE HELD May 16-18 at Rivier College in Nashua, NH. It opens with a Friday night hospitality suite followed by tech sessions and symposia all day Saturday and a noise figure and antenna gain session on Sunday. Dave Knight, KA1DT, 15Ø Oakdale, Nashua, NH Ø3Ø52 can provide details.



comments

fair comparison?

Dear HR

The article, "AMTOR, AX.25, and HERMES: A Performance Analysis of Three Systems" in the December, 1985, *ham radio*, purports to be a fair comparison between the three systems. However, on finding several errors in this article, and noting that the author, W9JD, is also the "inventor" of HERMES, I would like to take the opportunity, as the "inventor" of AMTOR, to correct these errors.

W9JD introduces the concept of "minimum required error-free time" being the minimum "clear channel" time required to get any traffic through at all, and he claims that AMTOR needs 4450 mS. This is not true: it is only necessary to have a 210-mS "clear patch" to get one block through; the corresponding "ack" need not return in the same clear patch but through any subsequent 70-mS clear patch. In this, the block-synchronous AMTOR system has an advantage over the other two systems. Even if, to make comparison easier, we insist that the "ack" must get through the same clear patch as the corresponding block, the minimum clear patch size is 280 mS, since the one signal follows immediately after the other in the same way as the other two systems.

In using the concept of "bit error-rate to stop" W9JD says that one bit error in every 450-mS AMTOR cycle will stop the link, deducing from this that the link will be stopped by a random B.E.R. of 2.2 percent (1 in 45). This is wrong on two counts. First, a random B.E.R. of 2.2 percent doesn't guarantee that every cycle will contain one error, and some will have none, allowing the link to continue to pass traffic. Second, since only 280 mS of the 450-mS cycle contain signals, a randomly-occurring error will "hit" only 28 out of every 45 cycles. In fact, a B.E.R. of 2.2 percent will only slow the link down to about 50 percent, not stop it.

In analyzing the probability of an undetected error in an AMTOR character, W9JD says that this will occur if the channel reverses an even number of bits. This is nonsense. An undetected error will occur only if the number of 0-to-1 corruptions equals the number of 1-to-0 corruptions. The probability of this occurring is highest with random noise input at 27 percent, not 50 percent. However, the probability of an undetected error in a block of three is higher than the third power of this when there is some signal in the noise. It reaches a peak of 5.6 percent when the B.E.R. is 11 percent, then falls rapidly to zero below this. (See IEEE's *Transactions on Communications*, July, 1985, page 710, for a full explanation of this strange effect.) However, 5.6 percent is still less than half of W9JD's figure of 12.5 percent.

While writing, I would like to introduce a concept which is very relevant when comparing systems for use in Amateur Radio, namely "latency," or the time taken for traffic to propagate through the system. Although not important for one-way broadcasts, excessive latency is a disadvantage in a two-way conversation, as anyone who has experienced the 240-mS delay of a satellite telephone link will testify! The latency of AMTOR is 0.21-0.45 second; of AX.25 (using figures from the W9JD article), 2.8-23.4 seconds; and of HERMES, 13.9-43.4 seconds. Its relevance to two-way QSOs is that it will take at least twice the latency figure to "pass it over" to the other station. Although the longer block sizes of AX.25 and HERMES make for lower undetected-error rates, their very high latency would, in my opinion, make them unusable for two-way conversational QSOs on HF, for which purpose AMTOR still leads the field.

J. Peter Martinez, G3PLX
Gosport, Hantshire, England

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by Doug DeMaw, W1FB

This new anthology has been taken from DeMaw's '84 and 85 series in QST magazine. It has been written to give beginners the basic electronic theory needed for upgrading and the ins and outs of how radios work. Using a building block approach, DeMaw first explains what the different componants are, then assembles them into basic circuits and ends up with how these circuits work in your radio. You also get articles on antennas, propagation and beginners level RFI problems and suggestions on how to resolve them. Great review for more experienced Hams. Perfect for the beginner. © 1985.

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by Edward Noll, W3FQJ

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See what high-performance stations do to achieve big-gun status on 80/160 meters

secrets of successful low band operation: part 1

I was curious . . . I would often hear stations on the low bands comfortably exchanging signal reports with rare DX contacts and wonder what type of antenna system they were using to achieve such good results. Therein was the beginning of a project that has provided more answers and technical enjoyment than I would ever have imagined possible.

Late last year a three-page survey was composed and sent to 113 of the "big gun" station operators and owners. After reading and digesting a two-inch thick sheaf of replies, many of my questions — as well as many unasked ones — have been answered. It is our pleasure to share this information with you. Much can be learned from the thousands of hours of experience in not only system design but also propagation observations made through the use of these outstanding systems.

A sincere note of thanks is due to the 47 individuals who took the time to fill out the lengthy survey form. Although a considerable amount of information was requested, some respondents not only entered their answers to questions but added more data as well. Responses were received from 21 countries from all continents. A special note of thanks goes to our overseas friends who, besides providing the data, responded in English, which is, for the most part, not their primary language. Responses were received from the following:*

CN2AQ	DJ0IA	DL0WU	EA8ADP	G2PU
G3WMZ	G4AMN	GW4OFQ	I5NPH	JA1FRE
JF1IST	K1MEM	K2FV	K3ZO	K5UR
KG7D	LA7ZO	N1ACH	N4AR	N4RJ
N4SU	NW5K	N6DKP	N7CKD	OE6MBG
OH1RY	OZ8BV	PA3DFU	SM4CAN	SM6EHY
SP3GEM	TI5EWL	VE2HQ	VE3BMV	VE7BS
VK6LK	W1FV	W1NH	WB2ITR	W2JB
W3BGN	W4DR	W6NLZ	W6RJ	YU7PFR
ZL4BO	4X4NJ			

a word on format

Originally, I had considered providing the information in a strictly tabular fashion, following the format of the survey with a list of callsigns and the data in the adjacent columns. But this approach would have required 13.5 printed pages, just for the data! Instead, each category is discussed and tabulated individually, with callsigns provided in some cases.

low band transmitting antennas

Wow! The variety and creativity evident in this area is outstanding and as diversified as the users themselves. One of the reasons people are attracted to the low bands is because it is (or at least was) virgin territory where you cannot simply buy your station and



fig. 1. Three full-size element homespun 80-meter Yagi helps produce the outstanding signal heard from I5NPH.

*A note to the others who received a copy of the survey, but whose names do not appear on the list: if you have not yet sent in the form please do so and the information will continue to be compiled. If a form was not sent to you please consider it an oversight on my part and send for it.

By Rich Rosen, K2RR, Editor-in-Chief, *ham radio*

Table 1. 80 meter transmitting antennas

Antenna type	D/P	Additional comments
Yagi, 4 el,tubing	P	Double driven, WoNLZ
Yagi, 3 el,tubing	P	Full size el,OH1RY,I5NPH
Yagi, 3 el,tubing	P	Loaded el, VE2HQ,P 32Ω,W6RJ
Yagi, 2 el,tubing	P	40% full size, N6DPF
Yagi, 3 el,wire	P	Wire added one side,YU7PFR
Half loops, 3 el,wire	P	Versatile system, N4SU
Delta Loop, two element	D	240 per loop,DJ0IA
Delta Loop, single element	-	@135 ,VE3BMV;@20m SF3GEM
Bobtail curtain, phased 2 bay	D	Vt el=238/freq, N4AR
Bobtail curtain, single bay	D	Remote-tuning cap, SM4CAN
Vertical arrays, 4 elements	D	W1NH W4DR K2BT SM6EHY
Vertical arrays, 3 elements	D	0.125 wave, 2 active,W1FV
Vertical arrays, 2 elements	D	HyTowers @ 0.5 wave,W3BGN
Vertical, quarter-wave	-	63 of Rohn 25, NW5K
Inverted Vee	-	@36 ,70 ,OE6MBG
Slopers, half-wave, 3 el	P	1 driven, 2 reflect,LA7ZO
Slopers, half-wave, 2 el	P	Fed 70° apart,WB2ITR
Slopers, half-wave, single	-	@80 , VK6LK
Slopers, quarter-wave	-	@100 ,K5UR
Half Square	D	Fed RG58 upper corner,K2FV
Dipoles, phased	D	Phase 0.25 wave both,K2FV
Dipole,tubing	-	94 rotary,@160 N4RJ
Dipole,wire	-	@50 ,GW4OFD
W8JK	D	140° el, @ 65 ,W2JB

Table 2. 160 meter transmitting antennas

Antenna type	D/P	Additional comments
Vertical plus reflector	P	Elec rotatable, 4x4NJ
Vertical, plus parasitic el	D	Cardioid, SM6EHY
Vertical, quarter-wave	-	Grounded monopole,N1ACH
Vertical, quarter-wave	-	Loaded, DL0WU
Half-wave loops, phased	P	Refl/Dir switching,N4SU
Inverted Vee	-	4" open wire @90 ,W2JB
Inverted L, phased	D	Hytower with 10B H,W3BGN
Inverted L	-	75 Vert,55 H, VE7BS
Slopers,quarter-wave,phased	P	Driven+3 refl,N4RJ
Slopers, quarter-wave	-	3 switched, tG7D
Sloping delta loop	-	Fed 26 fm corner,W3BGN

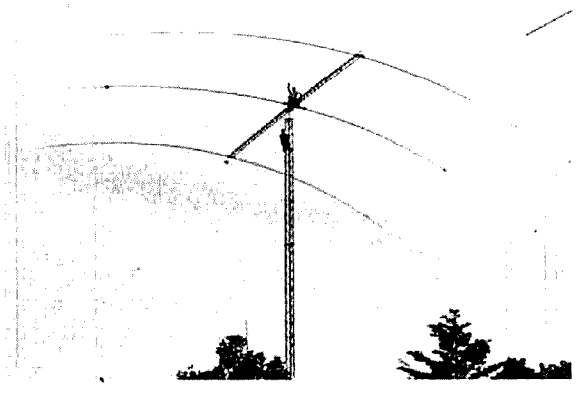


fig. 2. JF1IST uses a commercially available model CY-703 (Create Manufacturing) shortened three-element Yagi.

have someone else erect and install a competitive system. It meant — and still often means — climbing to greater heights, literally, and getting dirt under your fingernails. That's changed to some extent with the availability of commercial Yagis for 80 meters, but the challenge and achievement remain. The full-sized homebrew three-element 80-meter Yagi of I5NPH is seen on its way up (**fig. 1**). There is nothing "amateur" about this installation.

Tables 1 and 2 list the 80- and 160- meter Driven and Parasitic transmitting antennas used by a representative sample of those surveyed.

A few notable examples of some very effective 80-meter Yagi installations are apparent in **figs. 2, 3, 4,** and **5.**

low band receiving antennas

As a general rule, those surveyed agreed that two basic types of antennas are used for reception: the elaborate transmitting array or Beverages. That poor, often-repeated statement about verticals working (and hearing in this case) equally poorly in all directions is incorrect. As part of an array, vertical antennas exhibit directivity and discrimination against unwanted noise and QRM.

Beverages, for those who have the space, can also be used very effectively to receive signals from a preferred direction and discriminate against others. A two-wire version, recently known as an SWA and described by Beverage in his original AIEE paper[1] has the additional ability to rotate an azimuthal null. As with any other antenna, Beverages have their supporters and detractors. That they work is indisputable; that most are installed far from optimally is probably also correct.

Table 3 includes details of Beverages used effectively by the surveyed stations. Lengths, terminations, installation heights, preamp use, and callsigns are provided. Though most Beverages preferably are installed over low conductivity soil, OZ8BV proves that they still perform over salt water with his four unterminated 200-meter long antennas.

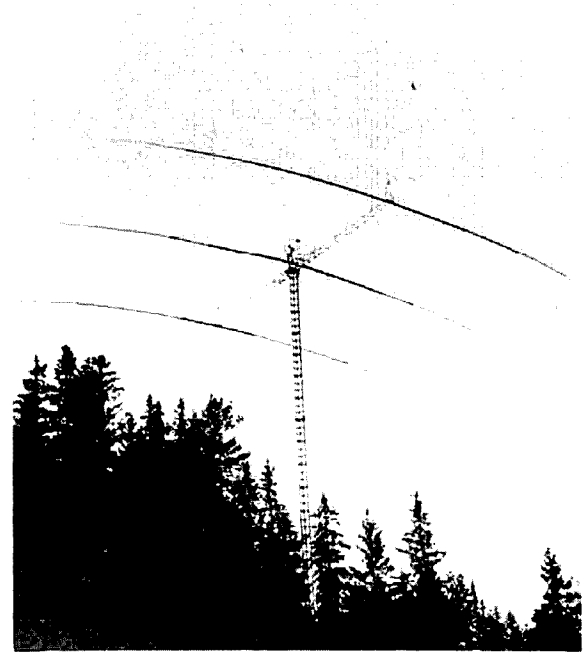


fig. 3. Full size element Yagi is installed "in the clear" and accounts for OH1RY's strong signal from Finland.

Table 3. Low band Beverage receiving antennas

Lengths		Term.	Notes	Callsign
Three	@300-600	Unterm		VE3BMV
Five	@540'	450 ohms	20dB preamp	N1ACH
One	@800'		10	4X4NJ
One	@390	440 ohms	8'	VK6LK
One	@350'	Bidirect	10 ,2 #19	VE7BS
Two	@750' One @500			W3BGN
Six	@1000			W1NH
Seven	@550	150-1100	3-12	K5UR
One			2 wire SWA	SM4CAN
Two	@700		2 wire	N4RJ
One	@500		1 wire	N4RJ
One			2 wire SWA	N7CNB
Four	@600		Reversible	N4SU
Three	@500		Fed both ends	W4DR
One	@500	Unterm	6 +preamp	K1MEM
One	@950'		8	K1MEM
One	@450	450 ohms		W1FV
One	@550	Unterm		W1FV

Table 4. Yagi material description

Nr of el.	Element description	Wgt.
3 (K3ZO)	90 ,linearly loaded (K1M)	400#
5 (OH1RY)	142' (R),135.5' (DR),129.7' (D),5 diff.dia. R-DR=37.7' , DR-D=34.45' (Homebrew)	1057#
2 (N6DKP)	El 40% full size, DR-R=24' (Homebrew)	
4 (W6NLZ)	90' el, 76' boom (K1M)	
3 (JF1IST)	96.5' max el,50' boom, (Create CY-703)	264#
3 (VE2HQ)	110' lin loaded el, 11 diff dia,65' boom	515#
2 (JA1FRE)	33' boom (Homebrew)	220#
3 (W6RJ)	(K1M)	

Table 5. VSWR bandwidths for typical 80/160 meter antennas

Antenna	VSWR	Bandwidth (kHz)	
3 el Yagi,shorter el	2:1	100 on CW,100 on SSB	(80)
3 el Yagi,full size	2:1	380	(80)
3 el Yagi,wire	2:1	240	(80)
2 el Yagi,40% full size	2:1	60	(80)
Dipole,rotary (K1M)	2:1	100-120	(80)
Vertical,4 el driven	2:1	300	(80)
Vertical,2 el driven	2:1	70	(160)
Vertical,130 shunt fed	2:1	40	(160)
Vertical,quarter-wave	2:1	100	(160)
Vertical,quarter-wave	2:1	375	(80)
Half-wave slopers	2:1	250	(80)
Delta Loop	2:1	75	(160)

antenna materials

This broad category discusses the materials used to assemble the previously described transmitting antennas. In general the Yagis use tapered sections of aluminum, if made from tubing, and copper-clad wire otherwise. The verticals using aluminum (6061-T6) or steel masts, or single or multiple bare or covered wires, show more variety. In addition, most of the rotatable Yagis use a combination of aluminum tubing and insulating sections that are linearly loaded by multiple turns of wire. **Table** 4 provides data on the rotaries in terms of length of elements, total weight of antenna and miscellaneous details. All of these are for 80 meter antennas. So far, to my knowledge, no one has an operational 160-meter rotatable Yagi.

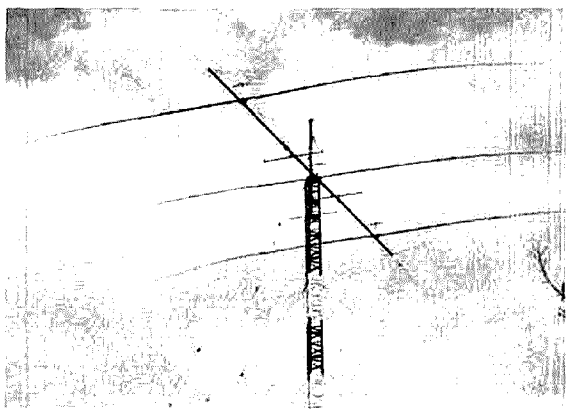


fig. 4. VE2HQ's homemade Yagi uses linearly-loaded elements but still is quite large.

hardware

This category encompasses the large rotaries, verticals and wire antennas. Once again the consensus is to use hardware that will last under all weather conditions. For the Yagis, this implies stainless steel or other non-rusting metals; for the wire antennas, good quality rope that is UV-resistant or "Phillystran." To withstand all weather conditions, liberal use of paint, tape, epoxy, Dow plastic sealant, Copper-kote™, or Coax-Seal™ is recommended where applicable. Remember that if the antenna is going to fail, it will most probably fail in the worst weather, during the low-banders' best operating season.

electrical characteristics

The great variety of responses to this question were not only a function of the particular antenna in use but also a matter of personal preference for instantaneous bandwidth operation versus tuning with a matching unit. There were those who wanted only flat 50-ohm lines using coax, and there were quite a few who thought that open wire line and use of a matchbox was a better approach. N4SU summed up the latter opinion with the following statement: "It boggles my mind to think about all the Megawatts of RF energy being wasted in heating coaxial cable on the ham bands in this country." An example of N4SU's approach is shown in **fig. 6** where he uses his multiloop array for 20, 40, 80 and 160 meters.

By my way of thinking, both approaches have their obvious advantages and disadvantages. **Table 5** summarizes some of the achievable VSWR bandwidths correlated with the particular antennas used. This is just a rough guide; VSWR is a function of many parameters.

antenna gain

Another somewhat controversial term is achievable gain. However, unlike the low-banders' higher frequency cousins (with their 10 to 20-meter high-gain arrays), due to the large element dimensions on 80 and 160, the actual variation between lower and higher gain antennas is not too great. After reviewing all the data, the highest gain antenna described, I believe, is the phased Bobtail curtains built by N4AR; he estimated the gain at 7 to 8 dB over a single ground-mounted vertical. The three-element Yagis are not far behind, with those surveyed indicating a range of 5.5 to 6 dB. Next in the gain line are the phased four-

element vertical arrays, coming in at between 4.9 and 6 dB, depending upon whom you talk to.

The textbooks provide, of course, the theoretical maximum values, but there are several important points that should be stressed before getting carried away with these numbers. Though the maximum difference in antenna gain between a *big gun* and a *little pistol* on 80 and 160 is around 6 dB, it is not the most important factor in the success story. The true criterion that makes these high performance stations shine is in their ability to *hear* — and I'm not necessarily talking about the use of Beverages.

A four-square vertical array does not have much more gain than a standard (quarter wave electrically and spatially separated) two-element array. But look at their front to back ratios, i.e., their E and H field patterns). The larger system provides considerably more attenuation off the back over a greater azimuthal beamwidth than does the cardioid version. This translates to receiving better and not disturbing others on adjacent frequencies as much if they're not in the beam direction when you're transmitting. It is for this reason, and the fact that it is difficult to quantify the received survey data (without all the parameters and operating conditions being known) that more specific antenna gains are not listed.

Before we leave this subject, that "measly" 6 dB variation in gain mentioned before can, at times, on the low bands, represent an enormous difference. During marginal conditions, even a 1 dB change in signal level can mean the difference between contact and no contact.

front-to-back ratios

Getting down to basics, what determines a high front-to-back ratio? "Front" is where the signals add and "back" is where they cancel, vectorially speaking. Even simple arrays (two elements) can experience a front-to-back (ratio) in excess of 60 dB! However, this is for very specific signal arrival angles (azimuthal and elevation) and polarization on sky-wave signals in particular. A slight change in one angle is all that is needed to upset the relationship. And changes do occur, sometimes over a very short period of time.

More complex arrays, however, can and are designed to produce azimuthal patterns that exhibit deep nulls (30 dB or more) over a 90-plus degree beamwidth. The four-square and the three-element inline are just two examples. The proponents of these driven arrays are quick to point out that they have two major advantages over rotary Yagis: instant direction change (through switching) and better front-to-back and front-to-side values. However, in fairness to the Yagi constituency, there are "Horizontal" nights and "Vertical" nights, and many a time a high horizontal beam has handily beaten the vertical arrays into Europe on the short path. But let's see, in **table 6**, what F/B's have been observed by the actual users of the low band antennas.

polarization

There are no eye-openers here. Both horizontally and vertically polarized antennas are used effectively by the high-performance stations. Due to the large antenna dimensions at these low frequencies, there are more vertical than horizontal *arrays*. However, as pointed out before, there is no one *best* antenna for all propagation conditions, locations, and times. If there were, it would probably have the following properties: instantaneous switching in azimuth, ele-

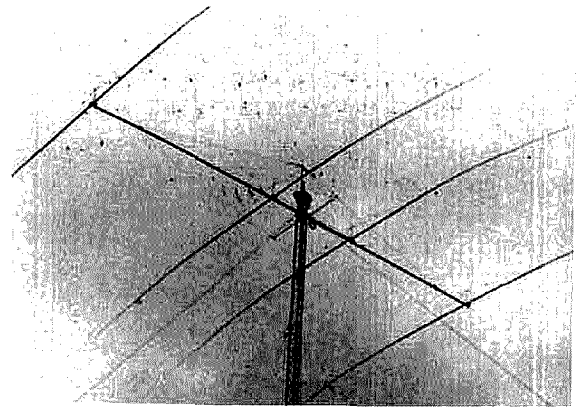


fig. 5. A double-driven element is used in W6NLZ's four-element Yagi, a product of KLM.

Table 6. F/B ratios for low band antennas

Antenna type	F/B (dB)	Comments
2 element Yagi,short el	10-15	JA1FRE
3 element Yagi,short el	10-15	function of freq.
3 element Yagi,full size	30	
3 element Yagi,wire	8-12	YU7PFR design
4 element Vert. phased array	30	W1NH
4 element Vert. phased array	15-45	SM6DHY
2 element Vert. phased array	20-30+	W1FV
Bobtail curtain, two bay	15-20	Great F/S
Delta loop, 2 element driven	20	DJ0IA

Table 7. Rotators

Designation	Comments	# in use
Telrex	#BA32B99R1HS (Largest)	3
Homebuilt	1:250 ratio, 0.5 rpm	1
Create RC5A-3X3	Manufactured in Japan	2
Modified prop-pitch	Rotates KLM 4 el Yagi	1

Table 8. Support structures

Description	Height	Station
Building plus tower	180	EA8BDF
Rohn 45	160	N4RJ
Rohn 80 plus 10' mast	150	K5ZO
Self-supporting	131	JA1FRE
Rohn 45G plus 10' mast	130	K5UR
Trees	130	VE7BS
Westover series tHD	120	LA7ZO
Tri-ex T-20	120	W6NLZ
Homebuilt 50' on 60' microwave twr	110	VE2HO
Telrex Big Bertha	110	VE3BMV
Heights	98	N4SU
Create K1-11N	98	JF1IST
Tri-ex Skyneedle	90	W6RJ
Trees	90	K2FV
Rohn 25	78	K1MEM
Pole	65	DJ0IA
Buildings	59	YU7PFR
Self-supporting	40	TI5EWL

vation and polarization, variable beamwidth control, and no sidelobes. In addition, it would have the capability of being scanned in wider arcs while simultaneously being operated in narrow mode. After daydreaming for a second, even the super stations compromise in this respect. That's basically why many use multiple antennas that offer different polarization and optimum angle of arrival reception capability.

steering

There are two basic means of steering low band arrays: mechanical and electrical. Two or more elements can be made to transmit or intercept signals from specific directions if certain amplitude and phase relationships at their terminals are satisfied. This is accomplished through the use of delay lines, lumped components, or combinations thereof. YU7PFR illustrates in **fig 7** his method of "steering" a three-element 80-meter wire array. The specific details have been provided in many articles and books. An excellent treatise on the design and construction process for phased vertical arrays (also applicable to horizontal arrays) can be found in a recent series of articles by Forrest Gehrke, K2BT.[2]

Mechanical rotation of the large arrays is not a task to be taken lightly. It requires careful consideration of many factors, including antenna wind-swept area,

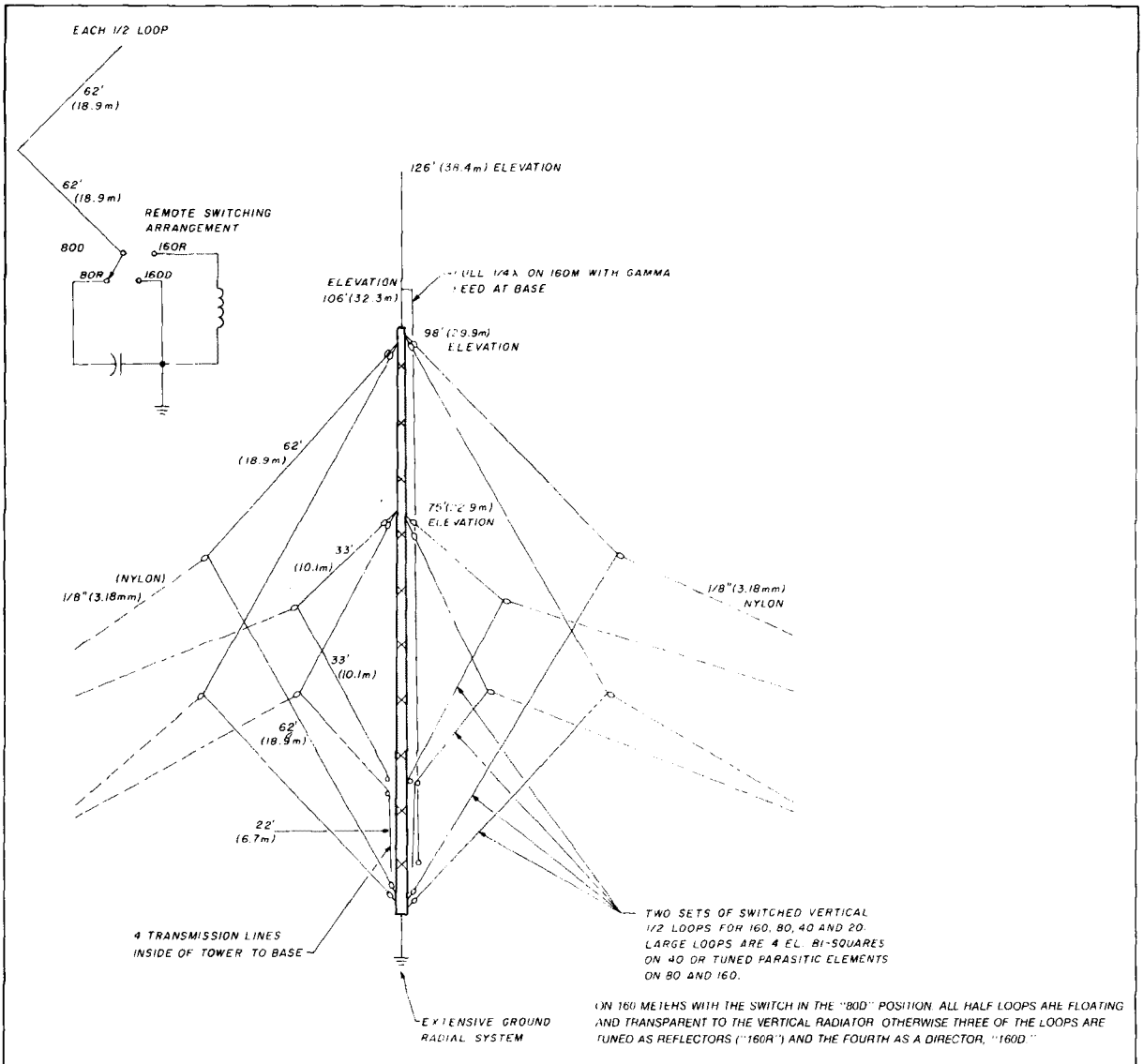


fig. 6. N4SU's multiband antenna array uses open wire line from the elements to the matching units.

mounting height, height above local terrain (i.e., is it shielded by trees from high winds?), weight, mast length, metallurgy, speed of rotation, weather conditions — both average and extreme — and expense, to name a few. To accomplish this task, the rotators listed in **table 7** are used by several of the superstations. Unfortunately, only a few of these responded to this particular survey question.

skyhooks that support these antennas

Man-made and natural structures consisting of towers, masts, buildings, and trees provide the support for the 80- and 160-meter antennas used by the stations surveyed. Considering the heights required, they probably represent the single greatest installation expense (excluding trees, of course). **Figure 8** illustrates the turnbuckles and guy lines used to tie down a 160-foot tower.

Towers and masts can and often do serve dual purpose as support structure and radiator, as in the case of verticals. The variety of the towers is considerable, with some installations surpassing, in quality, even those used by the commercial radio services.

Many not wishing to avail themselves of commercial units choose to build or modify other existing structures to suit their needs. Those fortunate enough to have tall trees on their property have been able to construct low band wire antennas that compete effectively with tower-mounted aluminum behemoths. On the other hand, it might surprise a few to see how *low* some of the big-gun stations have their antennas. **Table 8** is a compilation of some of the support structures used by the high-performance stations.

When a tower or mast is mounted on the roof of a building, the effective height of the antenna is not necessarily the combined heights. The building roof, depending on its dimensions, electrical characteristics, and separation from the antenna could determine an array factor with a much *higher* takeoff angle than would be indicated by the total height of the antenna above ground. In the case of EA8ADP, his strong signal into the United States would indicate that everything is working in his favor.

ground systems

This perennial question is asked all the time: how large a ground system is needed? The simple answer is *the larger the better*. Unless your antennas are situ-

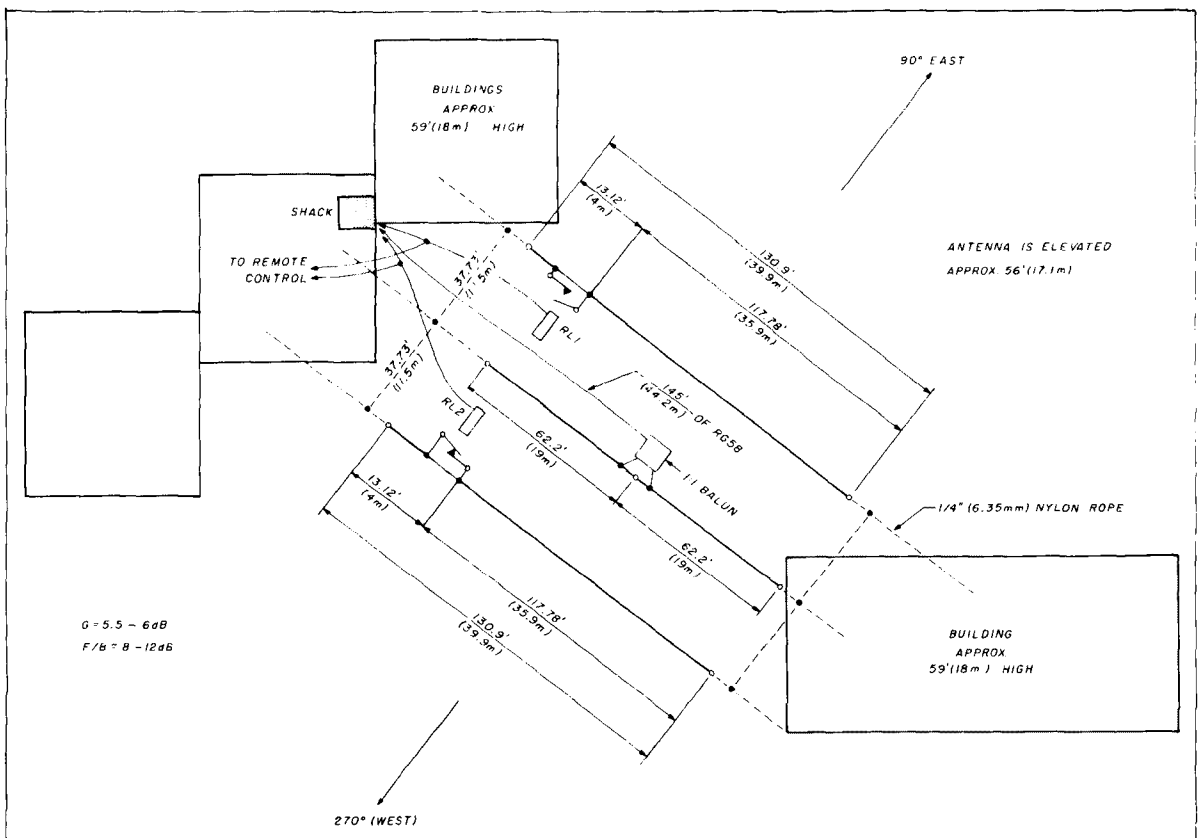


fig. 7. YU7PFR is able to switch antenna firing directions by opening and closing relays.

ated above an infinite extent, infinite conductivity ground plane, there is room for improvement. For those using verticals, two conditions should be met:
1. The immediate area around each vertical should have a high density of wire within a quarter wavelength of the radiators. A minimum of 100 radials is recommended. This determines and stabilizes the zone impedance and reduces losses, which is especially important if the current loop (current maximum) is close to or at ground level.
2. An extended ground or radial system should be installed as long as possible in the preferred transmission and reception directions. This increases the amount of energy present at truly low angles. *How long*, you ask? That's a function of several parameters, including propagation path and mode, distance, far field conductivity and permittivity (dielectric constant), solar activity, state of the geomagnetic field, transmitter power, and receive site conditions, to name just a few. Specifically, if I recall correctly, a six-wavelength radial system will produce a maximum elevation lobe at 6 degrees from the horizon using a single vertical.

SM6EHY suggested that it might be worthwhile to have the capability of remotely switching an extended radial ground system in and out. When switched in, lower angle of arrival signals are enhanced. When switched out, higher angle takes over. This could be accomplished through the use of convenient ground-bus point located relays. (Possibly another way to achieve this result would be through the use of *two* concentric ground rings at the antenna — the longer radials being attached to the outer switched in/out ring.) One useful application of enhanced lower angle transmission would be to "sneak under" the auroral high absorption layers. (More on this later under "propagation.")

Those using vertical antenna systems where the current loop is *not* at ground level should still be concerned with providing as large an extended radial system as possible for the second reason, even though, the ground loss resistance term represents a smaller fraction of the total antenna, in this case feed impedance.

A useful compromise for all vertical users is to place a mesh under the antenna as well as a radial system. The larger, of course, the better — but since maximum return currents *want* to exist within the immediate vicinity of the antenna, that's where it's most useful. The mesh, in addition to a 0.25 or 0.3 wavelength radial system, would improve the performance of the composite antenna system.

The existence of an extensive and symmetrical ground system also aids the phased vertical array designer. It eliminates one of the unknowns, or, stated differently, doesn't introduce yet another complex term to deal with. (Warning: before you begin quoting the above statements as gospel, I should mention that the information represents my own educated opinion, based on the work of others, whom, I believe, are correct).

An extensive ground system also aids the horizontal antenna user. It helps establish the distance between the antenna and its image (Green's function) and consequently determines the composite elevation pattern. This represents an increase of up to 6 dB in the total signal (sky and ground reflected wave) at a specific takeoff angle. If this angle is optimum for the particular path, then, in simple terms your signal will be louder at the point of reception and vice-versa. **Table 9** is a compilation of the ground systems used by the super-stations. Notice, however, that only a few of the stations that utilize horizontal arrays have extensive ground systems.

In addition, NW5K uses 30 square feet of chicken mesh; 4X4NJ puts 2 pounds of copper sulphate around his ground stakes, and N7CKD, besides using 55 square feet of chicken mesh and a half-mile of 5-foot (1.52 meters) chain link fence and barbed wire, pours 500 pounds of rock salt and 150 pounds of copper sulphate into trenches that he keeps moist at all times. (I'd be a wee bit concerned about the latter chemical, especially with regard to the possibility of its leaching into the water table).

I personally try to practice what I preach and am presently using 200 radials that vary in length from 65 to 300 feet (20 to 91 meters) in addition to an approximately 1000 square foot (93 square meter) ground mesh. And that's for just one vertical.

soil characteristics

As mentioned previously (see "ground systems"), the importance of a good ground, especially for vertical antenna users, cannot be stressed enough. There are those who are fortunate to have an antenna site whose soil has high conductivity or, even better, a high salt water table close to the surface. Under these circumstances the requirements for both near and far field enhancement are approached. An almost ideal site would consist of a high tower mounted antenna overlooking salt water on all sides.

Notice that vertically polarized antennas have been stressed. Though it is true that horizontally-polarized antennas are affected by a good ground system in the near field, it's the far field conditions of a horizontally-polarized antenna that show striking dissimilarities to that of a vertically-polarized antenna.

A figure of merit can be assigned to the reflective "nature" of the earth's surface in the form of a complex quantity that has both amplitude and phase terms. This reflection coefficient is very different for signals impinging on the earth that are vertically, rather than

Table 9. Ground systems

Radials	Length	Gnd rods	Length	Station
3	125'	3	8'	K3ZO
-	-	3	8'	VE3BMV
2	131'	-	-	PA3DFU
3	135'	4	10'	4X4NJ
65	65'+	-	-	SP3GEM
50	62'	-	-	VK6LK
20	75'	-	-	VE7BS
125	65'	4	8	NW5K
90	80'	6	6	W3BGN
-	-	9	6.6	JF1IST
320	80'	500	4	W1NH
100	130'	1	2	K5UR
65	130'	-	-	N4RJ
5	250'	-	-	N4RJ
400	70-1000'	30	10'	SM6EHY
496	65	-	-	W4DR
15	65	3	6	DJ0IA
300-360	30-130	-	-	W1FV

horizontally, polarized (i.e., the E-field of the electromagnetic wave is vertical or horizontal by definition). The reflection coefficient is a function of the conductivity and dielectric constant of the earth, the frequency, and the angle by which the wave strikes the surface.[3]

Let's consider three sites and observe the amplitude and phase of the ground reflected component for both vertical and horizontal polarization. The first site is a typical New England location with low conductivity, rocky soil. The second is farm land — the pastoral, low hills, and rich soil typical of Dallas, Texas or Lincoln, Nebraska. The third site consists of an installation over salt water — for example, any coastal location right on the beach. Let's also examine two different angles of arrival (or takeoff); 15 degrees for the long path and 45 degrees for the short path to Europe from the east coast. **Table 10** summarizes the various conditions and results.

For those who wish to replicate the calculations, I used the following soil parameters:

Wavelength: 80 meters
Frequency: 3.75 MHz
Rocky soil: Conductivity: 2 mS
Dielectric constant: 14
Pastoral: Conductivity: 30 mS
Dielectric constant: 20
Salty: Conductivity: 4.64 S
Dielectric constant: 81

Table 10. Reflection coefficient vs soil type and takeoff angle

Soil	Angle	Polarization	Amplitude	Phase
Rocky	15	Horizontal	88% of max	[illegible] deg
Pastoral	15	Horizontal	97%	1.65
Salty	15	Horizontal	99.75%	0.14
Rocky	15	Vertical	17%	75.53
Pastoral	15	Vertical	62%	25.02
Salty	15	Vertical	96%	[illegible]
Rocky	45	Horizontal	72%	6.24
Pastoral	45	Horizontal	92%	4.45
Salty	45	Horizontal	99.75%	0.14
Rocky	45	Vertical	51%	1.44
Pastoral	45	Vertical	84%	0.85
Salty	45	Vertical	98.67%	0.7

interpretation of results

There are a few eye-openers here that help to explain observations made over the years. Did you ever wonder, for example, why stations using a relatively low dipole (50 or 60 feet/15.24 or 18.28 meters) on 80 meters would often be heard just as well as those using verticals working against a good ground screen? Compare the seventh and tenth lines in **table 10**; the ground reflected component of the signal from the low horizontal antenna is actually stronger than the vertical antenna (72 percent of maximum versus 51 percent). Though this was calculated for rocky soil conditions, examination of the other entries shows that the low horizontal beats the vertical for all soil types at a *45* degree takeoff angle.

The situation is even more pronounced for a 15-degree takeoff angle condition. Compare lines 1, 2, and 3 with lines 4, 5, and 6; notice that for every soil type, the high horizontal beats the vertical. However, a horizontal antenna produces a maximum elevation lobe of 15 degrees when it is one wavelength up (250 feet/76.2 meters). There are probably only a few low-banders (WA1EKV, for example) who, thanks to a high tower and local topography, have their antennas at this height. Consequently, for that low-angle, low-path shot, it's perhaps easy to construct

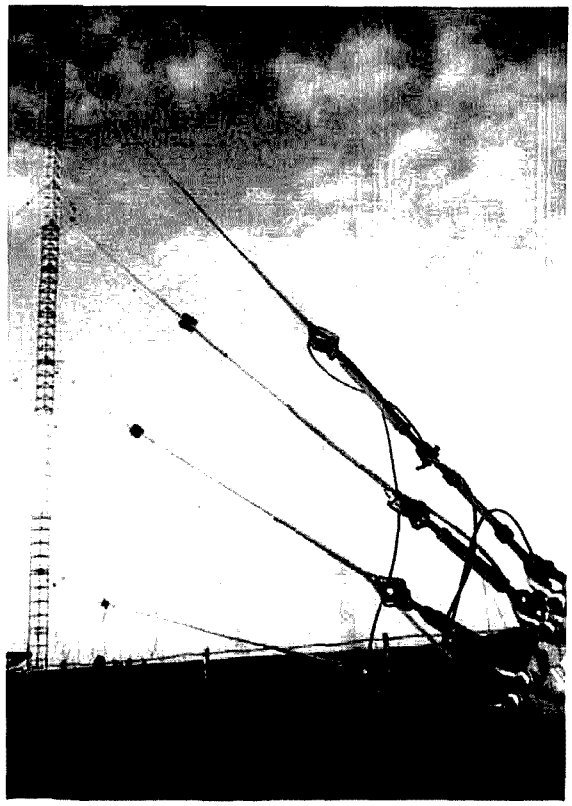


fig. 8. To gain an appreciation of the type of tower needed to handle low-band rotaries, examine VE2HQ's new 160-foot structure ready to support his three-element Yagi.

a vertical and get approximately the same performance as you would with the high horizontal. (This is true for medium to high conductivity far field soils).

Finally, if you're a lover of verticals, then the importance of an extended radial ground screen becomes apparent when you examine lines 4 and 6 in **table 10.** Copper, an even better conductor than salt water, provides an almost 600 percent improvement in reflected component level. This requires radials several wavelengths long to achieve a maximum elevation lobe of 15 degrees.

Let's say that you're considering several sites that have vastly different soil conditions for your ultimate super-station location. If you favor horizontal antennas, there's not a lot of difference in the reflected wave amplitude as a function of angle or soil parameters (compare lines 1 to 3 and 7 to 9 in **table 10.**)

But why are we so concerned with the reflected component? Simply because the composite launched wave or signal is a combination of the sky wave and the ground-reflected wave representing a doubling in signal level if everything works out. Which brings us to the next point: what factors are involved in maximum signal transmission/reception? The answer is that at the optimum launched angle (takeoff angle) — i.e. the angle best suited for that path, time of day, solar activity, and geomagnetic field conditions — the two components must add constructively. This means that the phase relationships must also match. Now, since the earth looks like a big capacitor, the reflected component always lags (negative phase angle) the incident wave. Notice in the last column of **table 10** the amount of phase delay is listed for the various soils, takeoff angles, and polarizations. It normally follows that if the amplitude of the reflected component is near unity, then the phase angle is quite small.

Upon examining all the survey responses no commonality between soil type and station performance could be discerned. The super-stations run the gamut from rocky to salt water ground characteristics and, in some cases, possibly have chosen their antennas carefully on this basis.

concluding installment

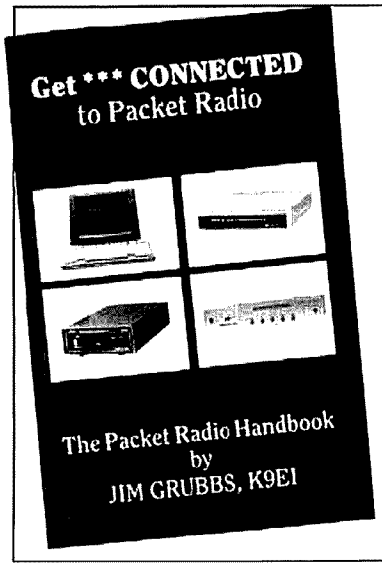
In Part 2 of this article, we'll examine responses from those surveyed to see how they rate their sites, including descriptions of near- and far-field conditions, obstructions, and noise sources with which they must contend.

We'll also consider the danger of lightning strikes and see what precautions these high-performance station operators, particularly those with exposed installations, have taken, both at the shack and at the antennas. Those who've sustained lightning strikes will describe the damage that occurred.

Construction of large antenna systems — whether they be rotaries, long wire antennas, or other elaborate systems — requires extensive planning, labor, expense, and maintenence. What periodic maintenence do these high-performance station operators recommend? Most of these stations have or are using different antennas — how do they compare? Next month, you'll also see how owners rate their stations against the competition. Propagation notes, including some startling results derived from thousands of hours of operating time by those with rotatable and switchable arrays, will be included.

It appears that the days of simple verticals and dipoles are rapidly passing; in our concluding installment, high-performance station owners and operators will describe additional improvements they plan to make to their stations to make them even more competitive in the future.

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computer-aided design of long VHF Yagi antennas

Design up to 40 elements on a single boom

This article describes a program for designing nine to 40-element Yagi antennas. You provide the design frequency, the number of elements, and boom and element diameters; the computer calculates the element lengths, element spacing, and several additional characteristics of the antenna. Whether the elements are insulated from or pass through the boom is also taken into consideration.

While this program breaks no new ground as far as antenna design is concerned, it does provide a quick way of designing long Yagis based on the designs of DL6WU, to whom all credit for these excellent antennas must be given.[1]

One of the most useful features of this program is the ability to easily change the number of elements and see — immediately — the resulting change in performance. You know, at once, whether the increase in length is worth the increase in gain obtained.

program description

The program was originally written on a TRS-80 Model 1 and given to a number of local VHF Amateurs, who were asked to evaluate the program and suggest improvements. This program is the result of that venture and resembles the original only slightly. The version printed here was written on an Apple IIe with a Z-80 processor, running Microsoft Basic™. Because this is a universal form of BASIC, there should be no difficulty using the program on most popular personal computers.

Three commands, peculiar to the computer on which this program was written, may need to be altered for use on other machines. They are **HOME**, to clear the screen and return the cursor; **INVERSE**, to change the background from light to dark, and **BEEP**, to sound a tone of predetermined pitch and duration. The latter two commands are cosmetic; no change in performance would result from their omission.

The disc-based program consists of the main program and 16 data files. (The data files could have been entered as data statements, but because only one is required each time the program is run, I opted for data files.)

The program begins in earnest at **line 160**, where the screen is cleared and the title screen is presented. (The code for this resides in **lines 1820** to **1920**.) After pressing any key, a few seconds will elapse while the computer reads the data for element spacing (**lines 180-220**), reflector multiplier (**lines 230-260**), radiator multiplier (**lines 270-300**), and element material size data (**lines 310-340**).

The next section (**lines 340-430**) asks for the design frequency, number of elements — which must be between 9 and 40 — the diameter of the boom and whether the elements are insulated or pass through the boom.

Lines 440-490 clear the screen and display the information entered. In **lines 500-640**, you're given the opportunity to change any of the input data.

Lines 650-790 calculate boom diameter in wavelengths, electrical boom length, beamwidths, and stacking distances. Data on the designed Yagi is displayed by **lines 800-940**. Should the design be unsatisfactory, **lines 950-990** allow you to start again.

Lines 1000-1100 input the size of the elements to be used. Calculation of the element lengths is achieved by lines **1120-1300**. Printing of the design and element data is done in **lines 1310-1750**.

operating the program

Operation is straightforward. When <RUN> is typed, the screen is cleared and the title is displayed. Pressing any key clears the screen, and after a few seconds you're asked to input "center frequency," "number of elements," "diameter of the boom," and whether "the elements are insulated from the boom." The screen is cleared and the input data is displayed. You'll

By David G. Hopkins, VK4ZF, #4 Handsworth Street, Capalaba, Queensland 4157, Australia

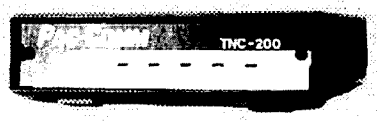
fig. 1. VK4ZF Yagi design program.

```
10 REM ************************************************************************
20 REM                          YAGI DESIGN PROGRAM.
30 REM                                BY
40 REM                         DAVID G HOPKINS  (VK4ZF)
50 REM                           # 4 HANDSWORTH ST
60 REM                                CAPALABA
70 REM                                QLD. 4157
80 REM                                AUSTRALIA
90 REM        PROGRAM BASED ON THE WORK DONE BY GUNTER HOCH (DL6WU)
100 REM                 AND PUBLISHED IN VHF. COMMUNICATIONS
110 REM                  REQUIRES THE USE OF LINE PRINTER.
120 REM                              V 2.2A
130 REM ************************************************************************
140 REM
150 REM
160 HOME : GOSUB 1820
170 CLEAR:DIM SP(40),DS(40),TS(40),LE(45):Q$="####":Q$(1)="###.##":Q$(2)="###.##
#":Q$(3)="#####.#"
180 REM----------------LOAD ELEMENT SPACING DATA-------------
190 DATA .240,.075,.180,.215,.250,.280,.300,.315,.330,.345,.360,.375,.385,.390,.
395,.400,.400,.400,.400,.400,.400,.400,.400,.400,.40
0,.400,.400,.400,.400,.400,.400,.400,.400,.400,.400,.400,.400,.400,.400
200 FOR X=1 TO 40
210 READ SP(X)
220 NEXT
230 REM--------------------LOAD REFLECTOR MULTIPLIER------------
240 DATA .4905,.4900,.4885,.4875,.4865,.4855,.4845,.4835,.4825,.4820,.4810,.4785
,.4770,.4765,.4750,.4740
250 DIM RE(17)
260 FOR X=1 TO 16:READ RE(X):NEXT
270 REM--------------------LOAD RADIATOR MULTIPLIER------------
280 DIM DR(17)
290 DATA .4675,.4665,.4640,.4620,.4601,.4585,.4575,.4550,.4530,.4515,.4500,.4460
,.4435,.4430,.4400,.4385
300 FOR X=1 TO 16:READ DR(X):NEXT
310 REM -------------------SELECT ELEMENT  MATERIAL SIZE----------------
320 DATA .003,.0035,.0042,.005,.0056,.0063,.007,.0078,.0088,.01,.01145,.0131,.01
5,.0165,.0182,.02
330 DIM EL(16):FOR X=1 TO 16:READ EL(X):NEXT
340 REM-----------------INPUT DESIRED PARAMETERS----------------
350 HOME :BEEP 20,10 :INPUT "WHAT IS THE CENTER FREQUENCY OF THE ANTENNA IN MHZ.
";F
360 BEEP 20,10:PRINT:INPUT"HOW MANY ELEMENTS DOES THE ANTENNA REQUIRE.";N
370 PRINT :IF N<9 OR N>40 THEN INVERSE :PRINT "NUMBER OF ELEMENTS MUST BE BETWEE
N  9 AND 40":NORMAL :GOTO 360
380 PRINT:BEEP 20,10:INPUT"WHAT IS THE DIAMETER OF THE BOOM IN MILLIMETERS.";BD
390 PRINT:BEEP 20,10:PRINT"ARE THE ELEMENTS TO BE INSULATED FROM THE BOOM Y=YES
N=NO ";
400 IN$=INKEY$ :IF LEN(IN$)=0 GOTO 400
410 IF IN$ ="Y" OR IN$= "y" THEN I=1: GOTO 450
420 IF IN$= "N" OR IN$="n" THEN I=0: GOTO 450
430 GOTO 400
440 REM -------------------DISPLAY SPEFICATIONS-------------------
  450 HOME :PRINT "SPEFICATIONS FOR THE ANTENNA TO DESIGN ":PRINT:PRINT:PRINT"1.
       DESIGN FREQUENCY :- ";F;" MHZ"
460 PPINT"2.           DIAMETER OF BOOM :- ";BD;" MILLIMETERS."
470 PPINT"3.           NUMBER OF ELEMENTS :- ";N
480 IF I=1 THEN PRINT"4.          ELEMENTS ARE TO BE INSULATED FROM THE BOOM"ELS
E PRINT"4.          ELEMENTS ARE TO BE NON INSULATED
 AND THPOUGH THE BOOM"
490 IF (BD/1000)/(299.792/F)>.05 THEN PRINT :PRINT :PRINT :PRINT "BOOM DIAMETER
IS TO LARGE FOR THIS FREQUENCY": PRINT :PRINT :PRINT
 "ENTER A SMALLER DIAMETER ":INPUT BD:GOTO 440
500 REM-------------------CHECK SPEFICATIONS----------------
510 PRINT :PPINT :PPINT :PRINT "ARE ALL THE ENTRIES CORRECT Y=YES  N=NO ";
520 IN$=INKEY$ :IF LEN(IN$) =0 THEN GOTO 520
530 IF IN$="Y" OR IN$="y" GOTO 650
540 IF IN$="N" OR IN$="n" GOTO 560
550 BEEP 100,10:GOTO 520
560 PRINT :INPUT "WHAT IS THE NUMBER OF THE INCORRECT ENTRY ";X
570 IF X< > 4 GOTO 600
580 IF I=0 THEN I=1: GOTO 440
590 IF I=1 THEN I=0: GOTO 440
600 INPUT "WHAT IS THE CORRECT VALUE ";CV
610 IF X=1 THEN F=CV
620 IF X=2 THEN BD=CV
630 IF X=3 THEN N=CV
640 GOTO 450
650 REM-------------CALCULATE BOOM DIAMETER WAVELENGTHS-------------
660 W=299.792/F
670 BW=BD/1000/W
680 BC=BD*1526.286*BW^.648831/100)
690 REM-------------------CALCULATE BOOM LENGTH--------------
700 TL=0
710 W=299.792/F
720 FOR X=1 TO N-1
730 DS(X)=SP(X)*W*1000
```


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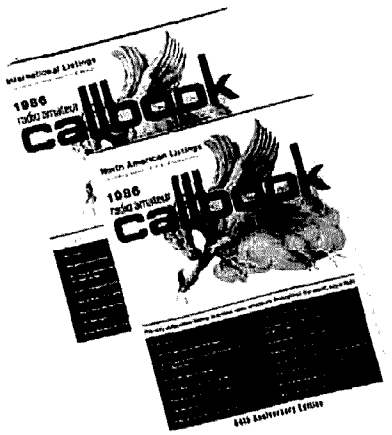
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```
740 TL=TL+DS(X)
750 NEXT
760 PG=(TL/1000)/W
770 REM-------CALCULATE BEAM WIDTHS AND STACKING DISTANCES------
780 BH=50.2709*PG^-.484091:SH=W/(2*(SIN(BH/2*.0174533)))
790 BV=66.5112*PG^-.617261:SV=W/(2*(SIN(BV/2*.0174533)))
800 REM-------------------PRINT PRELIMINARY DATA-------------
810 HOME
820 PRINT "          YAGI PRELIMINARY DATA":PRINT
830 PRINT "FREQUENCY                 ";F;" MHZ."
840 PRINT"WAVELENGTH                 ";W;" METERS."
850 PRINT "# OF ELEMENTS             ";N
860 PRINT "DIAMETER OF BOOM          ";BD;" MILLIMETERS"
870 PRINT "ELECTRICAL BOOMLENGTH     ";INT(TL);" MILLIMETERS"
880 PRINT "BOOM WAVELENGTHS          ";PG
890 PRINT "MAXIMUM PRACTICAL GAIN ";:PRINT USING G$(1);7.8*( LOG (PG)/ LOG (10))
+9.2;:PRINT " dBD"
900 PRINT "HORIZONTAL BEAM WIDTH   ";:PRINT USING G$(3);BH;:PRINT " DEGREES"
910 PRINT "VERTICAL BEAM WIDTH     ";:PRINT USING G$(3);BV;:PRINT " DEGREES"
920 PRINT "                STACKING DISTANCES"
930 PRINT "       HORIZONTAL:- ";:PRINT USING G$(2);SH;:PRINT "  METERS"
940 PRINT "       VERTICAL   :- ";:PRINT USING G$(2);SV;:PRINT "  METERS"
950 PRINT :PRINT :PRINT "DO YOU WISH TO CONTINUE WITH THIS DESIGN Y=YES  N=NO "I
960 IN$=INKEY$ :IF LEN(IN$) =0 GOTO 960
970 IF IN$="Y" OR IN$="y" GOTO 1000
980 IF IN$="N" OR IN$="n" GOTO 440
990 BEEP 100,10 :GOTO 960
1000 HOME :NU=0
1010 PRINT :PRINT "YOU MAY USE ANY OF THE FOLLOWING TUBING SIZES FOR THE ELEMENT
S"
1020 PRINT :PRINT "SELECT THE SIZE CLOSEST TO THE COMMERCIAL TUBE SIZE"
1030 FOR X=1 TO 16
1040 NU=NU+1
1050 PRINT "# ";X;".          ";:PRINT USING G$(3);(EL(X)*W)*1000;:PRINT " MM";:I
F (EL(X)*W)*1000>12 THEN X=16
1060 NEXT
1070 PRINT :PRINT :PRINT :PRINT "ENTER THE # OF THE TUBING SIZE YOU WISH TO USE"
;
1080 INPUT TS$
1090 IF VAL(TS$) >NU THEN BEEP 100,10 :GOTO 1000
1100 TS=VAL(TS$) :TT=EL(TS):TD(1)=TT*W*1000
1110 HOME :PRINT "STAND BY THIS WILL TAKE A FEW SECONDS"
1120 REM-----------------LOAD TUBING SIZE TABLES---------------
1130 N$="CURVE":N2$=N$+TS$
1140 OPEN "I",1,N2$
1150 FOR X=1 TO 38
1160 LINE INPUT #1,O$
1170 TS(X)=VAL(O$)
1180 NEXT
1190 CLOSE
1200 REM---------------CALCULATE ELEMENT LENGTHS---------------
1210 FOR X=3 TO 42
1220 IF I=1 THEN LE(X)=W*TS(X-2)*1000 ELSE LE(X)=(W*TS(X-2))*1000+BC
1230 NEXT
1240 LE(1)=W*RE(TS)*1000
1250 IF I=0 THEN LE(1)=LE(1)+BC
1260 LE(2)=W*DR(TS)*1000
1270 IF I=0 THEN LE(2)=LE(2)+BC
1280 HOME
1290 PRINT "PRESS ANY KEY WHEN THE PRINTER IS READY":BEEP 10,20
1300 IN$=INKEY$ :IF LEN(IN$) =0 GOTO 1300
1310 LPRINT "                          YAGI DESIGN DETAILS :-"
 1320 LPRINT "-------------------------------------------------------------------"
1330 LPRINT :LPRINT
1340 LPRINT "DESIGN FREQUENCY :- ";F;"  MHZ."
1350 LPRINT "WAVELENGTH       :- ";:LPRINT USING G$(2);W;:LPRINT "  METERS"
1360 LPRINT "NUMBER OF ELEMENTS :- ";N
1370 LPRINT "DIAMETER OF BOOM   :- ";BD;" MM"
1380 LPRINT "DIAMETER OF ELEMENTS :- ";: LPRINT USING G$(3);TD(1);: LPRINT " MM"
1390 IF I=1 THEN LPRINT "ELEMENTS ARE INSULATED FROM THE BOOM" ELSE LPRINT "ELEM
ENTS ARE NOT INSULATED FROM AND PASS THROUGH THE BOO
M"
1400 LPRINT "ELECTRICAL BOOM LENGTH :- ";INT(TL);"MM"
1410 LPRINT "BOOM WAVE LENGTHS :- ";PG
1420 LPRINT "MAXIMUM PRACTICAL GAIN :- ";: LPRINT USING G$(1);7.8*( LOG (PG)/ LO
G (10))+9.2;:LPRINT "  dBD"
1430 LPRINT "HORIZONTAL BEAM WIDTH :- ";:LPRINT USING G$(3);BH;:LPRINT " DEGREES
"
1440 LPRINT "VERTICAL BEAM WIDTH   :- ";:LPRINT USING G$(3);BV;:LPRINT " DEGREES
"
1450 LPRINT "HORIZONTAL STACKING DISTANCE :-";: LPRINT USING G$(2);SH;: LPRINT "
 METERS"
1460 LPRINT "VERTICAL STACKING DISTANCE :- ";: LPRINT USING G$(2);SV;: LPRINT "
METERS"
1470 LPRINT :LPRINT
1480 LPRINT "    ELEMENT LENGTHS IN MILLIMETERS.              DISTANCE FROM REFLE
CTOR MM."
1490 LPRINT "  ================================================================
============"
```

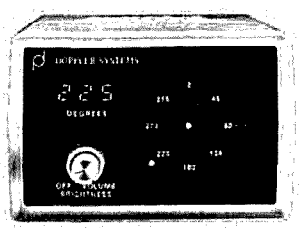
```
1500 LPRINT
1510 LPRINT "REFLECTOR          ";: LPRINT USING 0$(3);LE(1);: LPRINT ,: LPRINT ,:L
PRINT "  0"
1520 LPRINT "DRIVEN             ";: LPRINT USING 0$(3);LE(2);: LPRINT ,: LPRINT,:L
PRINT USING 0$(3); SP(1)*W#1000
1530 PS=SP(1)*W#1000
1540 FOR X=3 TO N
1550 PS=PS+SP(X-1)*W#1000
1560 LPRINT "DIRECTOR # ";X-2,: LPRINT USING 0$(3);LE(X),:LPRINT ,:LPRINT ,:LPRI
NT USING 0$(3);PS
1570 NEXT
1580 LPRINT
1590 LPRINT "    ELEMENT SPACING IN MILLIMETERS "
1600 LPRINT "======================================="
1610 LPRINT
1620 LPRINT "REFLECTOR - DRIVEN ",:LPRINT USING 0$(3);DS(1)
1630 LPRINT "DRIVEN    - DIR  1",:LPRINT USING 0$(3);DS(2)
1640 FOR X=3 TO N-1
1650 LPRINT "DIR";X-2;"     - DIR ";X-1,:LPRINT USING 0$(3);DS(X)
1660 NEXT
1670 LPRINT :LPRINT :LPRINT"                                        NOTES"
 1680 LPRINT :LPRINT "1.  THE DIMENSIONS ARE FROM CENTER TO CENTER IN ALL CASES"
1690 LPRINT "    FOR EXAMPLE THIS MEANS THE BOOM MUST BE CUT LONGER THAN THAT GI
VEN"
1700 LPRINT"    TO BE ABLE TO MOUNT THE ELEMENTS."
1710 LPRINT "2.  IF YOU WANT WIDE BANDWIDTH USE A FOLDED DIPOLE AS THE DRIVEN EL
EMENT."
1720 LPRINT "3.  THE DRIVEN ELEMENT DIMENSION IS THE LENGTH OVERALL."
 1730 LPRINT "4.  YOU MUST WORK TO AN ACCURACY OF BETTER THAN 1MM AT FREQUENCIES
ABOVE 400MHZ."
1740 LPRINT "5.  ACCURACY BELOW 400MHZ SHOULD BE WITHIN 1.5 MM."
1750 LPRINT "6.  ELEMENT MOUNTING MUST BE BETTER THAN .5 MM OF THE ELEMENT CENTE
R."
1760 HOME :FOR X= 1 TO 8:LPRINT :NEXT
1770 PRINT "ANOTHER CALCULATION ?  Y=YES    N=NO";
1780 IN$=INKEY$:IF LEN(IN$) =0 GOTO 1780
1790 IF IN$ ="Y" OR IN$="y" THEN GOTO 440
1800 IF IN$="N" OR IN$="n" THEN END
1810 BEEP 100,20 :GOTO 1780
1820 HOME :PRINT :PRINT "THIS PROGRAM WILL DESIGN LONG YAGI ANTENNAS"
1830 PRINT "   WITH ANY NUMBER OF ELEMENTS BETWEEN 6 AND 40"
1840 PRINT :PRINT :PRINT
1850 PRINT " THE PROGRAM IS BASED ON ARTICLES BY GUNTER HOCK"
1860 PRINT "  AND PUBLISHED IN VHF. COMMUNICATIONS"
1870 PRINT :PRINT :PRINT
1880 PRINT " YOU WILL REQUIRE A LINE PRINTER TO OBTAIN THE RESULTS"
1890 PRINT :PRINT :INVERSE :PRINT "PRESS ANY KEY TO CONTINUE ";: NORMAL:BEEP 20,
10
1900 IN$=INKEY$ : IF LEN (IN$)=0 GOTO 1900
1910 HOME
1920 RETURN
1930 END
```

be asked if you wish to correct any input. If all is correct the screen displays the information shown in **fig. 1**. Once again, you're asked whether you wish to continue or go back and change any input value.

If the design is acceptable, you next select the size of the tubing to use for the elements. Depending on the frequency, up to 16 different sizes of element material will be offered. Choose the size nearest the one you can readily buy from your supplier. The screen will now clear and after about 15 seconds, the computer will ask if the printer is ready, then print out the complete design.

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MRF421C	110W	—	60.00
MRF422*	150W	38.00	82.00
MRF426,/A*	25W	18.00	42.00
MRF428**	150W	55.00	125.00
MRF433	12.5W	12.00	30.00
MRF435*	150W	42.00	90.00
MRF449,/A	30W	12.50	30.00
MRF450,/A	50W	14.00	31.00
MRF453,/A	60W	15.00	35.00
MRF454,/A	80W	16.00	36.00
MRF455,/A	80W	12.00	28.00
MRF458	80W	20.00	46.00
MRF460	80W	18.00	42.00
MRF464*	80W	25.00	60.00
MRF466*	40W	18.75	48.00
MRF475	12W	3.00	9.00
MRF476	3W	2.75	8.00
MRF477	40W	11.00	25.00
MRF479	15W	10.00	23.00
MRF485*	15W	6.00	15.00
MRF492	90W	18.00	40.00
SRF2072	75W	15.00	33.00
SRF3662	110W	28.00	60.00
SRF3775	75W	15.50	34.00
SRF3795	85W	16.50	37.00
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SD1076	70W	17.00	40.00
SD1451	50W	15.00	36.00

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	Rating	MHz	Net Ea.	Match Pr.
MRF212	10W	136-174	$16.00	—
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MRF222	25W	136-174	14.00	—
MRF224	40W	136-174	13.50	32.00
MRF231	3.5W	66-88	10.00	—
MRF234	25W	66-88	15.00	39.00
MRF237	4W	136-174	3.00	—
MRF238	30W	136-174	12.00	—
MRF239	30W	136-174	15.00	—
MRF240	40W	136-174	18.00	—
MRF245	80W	136-174	28.00	65.00
MRF247	75W	136-174	27.00	63.00
MRF250	50W	27-174	20.00	46.00
MRF260	5W	136-174	7.00	—
MRF261	10W	136-174	9.00	—
MRF262	15W	136-174	9.00	—
MRF264	30W	136-174	13.00	—
MRF607	1.75W	136-174	3.00	—
MRF641	15W	407-512	22.00	—
MRF644	25W	407-512	24.00	54.00
MRF646	40W	407-512	26.50	59.00
MRF648	80W	407-512	33.00	69.00
2N3866*	1W	30-200	1.25	—
2N4427	1W	136-174	1.25	—
2N5591	25W	136-174	13.50	34.00
2N5842*	20W	30-200	13.75	34.50
2N5945	4W	407-512	10.00	—
2N5946	10W	407-512	12.00	—
2N8080	4W	136-174	6.25	—
2N8081	15W	136-174	7.50	—
2N6082	25W	136-174	8.90	—
2N6083	30W	136-174	9.30	24.00
2N6084	40W	136-174	11.75	28.50

TMOS FET

MRF134*	5W	2-200	$10.50	—
MRF137*	30W	2-200	22.50	—
MRF138**	30W	1.5-150	35.00	—
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```
        YAGI PRELIMINARY DATA
FREQUENCY                  432  MHZ.
WAVELENGTH                 .693963  METERS.
# OF ELEMENTS              15
DIAMETER OF BOOM           20  MILLIMETERS
ELECTRICAL BOOMLENGTH      2803  MILLIMETERS
BOOM WAVELENGTHS           4.04
MAXIMUM PRACTICAL GAIN     13.93 dBD
HORIZONTAL BEAM WIDTH      25  DEGREES
VERTICAL BEAM WIDTH        28  DEGREES
              STACKING DISTANCES
        HORIZONTAL:-    1.603  METERS
        VERTICAL  :-    1.434  METERS
```

fig. 2. Preliminary data display.

The first part of the printout provides the characteristics of the Yagi followed by the element lengths and the progressive distance from the reflector to each element. The next section gives the distance between the centers of the elements and concludes with some notes about the construction of the antenna.

The program is simple to use, in that any numerical input requires the pressing of the enter (return) key, but any input requiring a "Y" or "N" does not. Either upper or lower case may be used for the "Y" or "N" inputs. In the version shown, a *high* pitched tone is sounded when the computer requires an input and a *low* tone is sounded when it receives an unexpected input — i.e., when you've made a mistake. (This feature can easily be removed if your computer doesn't support the BEEP command.)

typing in the program

There's a lot of typing to do. *No errors are allowed.* The main program should pose no real problem, but the data files are a real chore. You may use the listing

```
10 HOME
20 DIM A(39):C$="CURVE"
30 PRINT "WHAT IS THE NUMBER OF THE CURVE YOU WISH TO ENTER ";:INPUT C$
40 F$=C$+".C"
50 PRINT "ENTER THE VALUES FOR THE CURVE"
60 FOR X=1 TO 38
70 PRINT"#";X;":";:INPUT A(X)
80 NEXT
90 HOME
100 PRINT "THESE ARE THE VALUE FOR CURVE";C
110 PRINT
120 PRINT
130 FOR X=1 TO 38
140 PRINT"#"; X;". ";A(X)
150 NEXT
160 PRINT
170 PRINT "ARE ALL THE VALUES CORRECT Y =YES   N=NO":INPUT A$
180 IF A$="Y" GOTO 250
190 IF A$="N" GOTO 210
200 GOTO 170
210 PRINT "WHICH # IS INCORRECT" :INPUT D
220 PRINT "WHAT IS THE CORRECT VALUE FOR :- #";D
230 INPUT A(D)
240 GOTO 180
250 HOME
260 PRINT "OUTPUT TO DISC"
270 OPEN "O" ,1, F$
280 FOR X= 1 TO 38
290 PRINT #1,A(X)
300 NEXT
310 CLOSE
320 PRINT "END OF PROGRAM"
330 END
```

fig. 4. Program to input data files.

```
                YAGI DESIGN DETAILS :-

DESIGN FREQUENCY :-          145.825   MHZ.
WAVELENGTH       :-            2.056  METERS
NUMBER OF ELEMENTS :-          9
DIAMETER OF BOOM   :-         25  MM
DIAMETER OF ELEMENTS :-       10.3 MM
ELEMENTS ARE NOT INSULATED FROM AND PASS THROUGH THE BOOM
ELECTRICAL BOOM LENGTH :-     3813 MM
BOOM WAVE LENGTHS :-          1.855
MAXIMUM PRACTICAL GAIN :-     11.29  dBD
HORIZONTAL BEAM WIDTH :-       32.3 DEGREES
VERTICAL BEAM WIDTH   :-       45.4 DEGREES
HORIZONTAL STACKING DISTANCE :-          3.216 METERS
VERTICAL STACKING DISTANCE :-            2.662 METERS

   ELEMENT LENGTHS IN MILLIMETERS.        DISTANCE FROM REFLECTOR MM.

REFLECTOR                  1089.7                      0
DRIVEN                      957.3                    493.4
DIRECTOR #  1               909.6                    847.6
DIRECTOR #  2               899.3                   1317.6
DIRECTOR #  3               889.5                   1939.6
DIRECTOR #  4               880.2                   1923.6
DIRECTOR #  5               871.0                   2549.2
DIRECTOR #  6               863.9                   3166.8
DIRECTOR #  7               857.0                   3813.6

   ELEMENT SPACING IN MILLIMETERS

REFLECTOR - DRIVEN          493.4
DRIVEN      DIR  1          154.2
DIR 1       DIR  2          370.1
DIR 2     - DIR  3          442.0
DIR 3     - DIR  4          514.0
DIR 4     - DIR  5          525.6
DIR 5       DIR  6          616.8
DIR 6     - DIR  7          647.6

                        NOTES

1.  THE DIMENSIONS ARE FROM CENTER TO CENTER IN ALL CASES
    FOR EXAMPLE THIS MEANS THE BOOM MUST BE CUT LONGER THAN THAT GIVEN
    TO BE ABLE TO MOUNT THE ELEMENTS.
2.  IF YOU WANT WIDE BANDWIDTH USE A FOLDED DIPOLE AS THE DRIVEN ELEMENT.
3.  THE DRIVEN ELEMENT DIMENSION IS THE LENGTH OVERALL.
4.  YOU MUST WORK TO AN ACCURACY OF BETTER THAN 1MM AT FREQUENCIES ABOVE 400MHZ.
5.  ACCURACY BELOW 400MHZ SHOULD BE WITHIN 1.5 MM.
6.  ELEMENT MOUNTING MUST BE BETTER THAN .5 MM OF THE ELEMENT CENTER.

                YAGI DESIGN DETAILS :-

DESIGN FREQUENCY :-          432.5   MHZ.
WAVELENGTH       :-            0.693  METERS
NUMBER OF ELEMENTS :-          20
DIAMETER OF BOOM   :-          19  MM
DIAMETER OF ELEMENTS :-         6.1 MM
ELEMENTS ARE NOT INSULATED FROM AND PASS THROUGH THE BOOM
ELECTRICAL BOOM LENGTH :-     4183 MM
BOOM WAVE LENGTHS :-          6.035
MAXIMUM PRACTICAL GAIN :-     15.29  dBD
HORIZONTAL BEAM WIDTH :-       21.1 DEGREES
VERTICAL BEAM WIDTH   :-       21.9 DEGREES
HORIZONTAL STACKING DISTANCE :-          1.897 METERS
VERTICAL STACKING DISTANCE :-            1.822 METERS

   ELEMENT LENGTHS IN MILLIMETERS.        DISTANCE FROM REFLECTOR MM.

REFLECTOR                   344.1                      0
DRIVEN                      323.7                    166.4
DIRECTOR #  1               306.3                    218.3
DIRECTOR #  2               302.4                    343.1
DIRECTOR #  3               298.8                    492.1
DIRECTOR #  4               295.1                    665.4
DIRECTOR #  5               291.0                    859.5
DIRECTOR #  6               286.8                   1067.5
DIRECTOR #  7               286.2                   1285.8
DIRECTOR #  8               283.6                   1514.6
DIRECTOR #  9               281.8                   1753.7
DIRECTOR #  10              280.0                   2003.2
DIRECTOR #  11              278.6                   2263.2
DIRECTOR #  12              277.1                   2530.8
DIRECTOR #  13              275.9                   2808.4
DIRECTOR #  14              274.8                   3074.2
DIRECTOR #  15              273.7                   3351.4
DIRECTOR #  16              272.6                   3628.7
DIRECTOR #  17              271.6                   3986.8
DIRECTOR #  18              270.7                   4183.2

   ELEMENT SPACING IN MILLIMETERS

REFLECTOR - DRIVEN          166.4
DRIVEN    - DIR  1           52.0
DIR 1     - DIR  2          124.8
DIR 2     - DIR  3          149.8
DIR 3     - DIR  4          173.3
DIR 4     - DIR  5          194.1
DIR 5     - DIR  6          207.9
DIR 6     - DIR  7          218.3
DIR 7     - DIR  8          229.7
DIR 8     - DIR  9          239.1
DIR 9     - DIR  10         249.5
DIR 10    - DIR  11         259.9
DIR 11    - DIR  12         266.9
DIR 12    - DIR  13         270.3
DIR 13    - DIR  14         273.8
DIR 14    - DIR  15         277.3
DIR 15    - DIR  16         277.3
DIR 16    - DIR  17         277.3
DIR 17    - DIR  18         277.3

                        NOTES

1.  THE DIMENSIONS ARE FROM CENTER TO CENTER IN ALL CASES
    FOR EXAMPLE THIS MEANS THE BOOM MUST BE CUT LONGER THAN THAT GIVEN
    TO BE ABLE TO MOUNT THE ELEMENTS.
2.  IF YOU WANT WIDE BANDWIDTH USE A FOLDED DIPOLE AS THE DRIVEN ELEMENT.
3.  THE DRIVEN ELEMENT DIMENSION IS THE LENGTH OVERALL.
4.  YOU MUST WORK TO AN ACCURACY OF BETTER THAN 1MM AT FREQUENCIES ABOVE 400MHZ.
5.  ACCURACY BELOW 400MHZ SHOULD BE WITHIN 1.5 MM.
6.  ELEMENT MOUNTING MUST BE BETTER THAN .5 MM OF THE ELEMENT CENTER.
```

fig. 3. Sample printouts.

CURVE1	CURVE2	CURVE3	CURVE4	CURVE5
.4478	.4448	.4418	.4388	.4366
.4433	.4401	.437	.4338	.4315
.4389	.4356	.4323	.429	.4267
.4354	.4318	.4281	.4245	.422
.4316	.4277	.4239	.42	.4175
.4281	.4242	.4204	.4165	.4138
.4251	.4211	.4172	.4132	.4105
.4222	.4181	.4141	.41	.4072
.4196	.4156	.4116	.4076	.4047
.4175	.4135	.4094	.4054	.4024
.4158	.4117	.4076	.4035	.4005
.4141	.41	.4058	.4017	.3986
.4125	.4083	.4042	.4	.3969
.4109	.4067	.4026	.3984	.3953
.4096	.4053	.4011	.3968	.3937
.4082	.4039	.3995	.3952	.3921
.4071	.4027	.3984	.394	.3908
.406	.4016	.3972	.3928	.3896
.4048	.4004	.3959	.3915	.3883
.4036	.3992	.3947	.3903	.3871
.4026	.3981	.3937	.3892	.386
.4017	.3972	.3926	.3881	.3849
.4009	.3963	.3918	.3872	.384
.4	.3954	.3908	.3862	.383
.3992	.3946	.39	.3854	.3821
.3985	.3939	.3892	.3846	.3813
.3979	.3932	.3884	.3837	.3804
.3971	.3924	.3876	.3829	.3796
.3965	.3917	.3869	.3821	.3788
.396	.3912	.3863	.3815	.3781
.3954	.3905	.3857	.3808	.3774
.3949	.39	.3851	.3802	.3768
.3943	.3894	.3845	.3796	.3763
.3938	.3889	.3839	.379	.3756
.3933	.3884	.3834	.3785	.3751
.3929	.3879	.383	.378	.3746
.3925	.3875	.3825	.3775	.3741
.3922	.3871	.3821	.377	.3736

CURVE6	CURVE7	CURVE8	CURVE9	CURVE10
.4345	.4323	.4301	.4279	.4257
.4293	.427	.4247	.4223	.42
.4243	.422	.4195	.4171	.4146
.4196	.4171	.4144	.4118	.4091
.4151	.4126	.4098	.407	.4042
.4112	.4085	.4058	.4027	.3998
.4077	.405	.4019	.3989	.3958
.4043	.4015	.3983	.3952	.392
.4019	.399	.3957	.3925	.3892
.3995	.3965	.3933	.39	.3868
.3974	.3944	.3911	.3879	.3846
.3955	.3924	.3891	.3858	.3825
.3937	.3906	.3873	.3841	.3808
.3921	.389	.3857	.3825	.3792
.3905	.3874	.3841	.3809	.3776
.3889	.3858	.3826	.3793	.1761
.3876	.3844	.3812	.3779	.3747
.3864	.3832	.3799	.3766	.1733
.3852	.382	.3786	.3753	.1719
.384	.3808	.3774	.374	.1706
.3829	.3797	.3763	.3728	.1694
.3817	.3785	.3751	.3716	.3682
.3807	.3775	.374	.3706	.1671
.3797	.3765	.373	.3695	.166
.3789	.3756	.3721	.3685	.165
.378	.3747	.3711	.3676	.364
.3772	.3739	.3703	.3666	.163
.3763	.373	.3693	.3657	.162
.3754	.3721	.3685	.3648	.1612
.3747	.3713	.3677	.364	.1604
.374	.3706	.367	.3633	.3597
.3734	.37	.3664	.3627	.1591
.3729	.3696	.3658	.3621	.3583
.3722	.3688	.3651	.3613	.3576
.3716	.3682	.3645	.3607	.157
.3711	.3677	.364	.3602	.1565
.3706	.3672	.3635	.3597	.156
.3702	.3668	.363	.3592	.1554

CURVE11	CURVE12	CURVE13	CURVE14	CURVE15	CURVE16
.4231	.4206	.418	.4153	.4127	.41
.4173	.4146	.4119	.4089	.406	.403
.4117	.4089	.406	.4028	.3997	.3965
.4062	.4034	.4005	.3973	.3941	.3909
.4011	.3981	.395	.3918	.3885	.3853
.3965	.3933	.39	.3869	.3838	.3807
.3926	.3894	.3862	.383	.3798	.3766
.3888	.3857	.3825	.3792	.3758	.3725
.386	.3828	.3796	.3762	.3728	.3694
.3835	.3803	.377	.3736	.3703	.3669
.3813	.378	.3747	.3713	.368	.3646
.3792	.3758	.3725	.3691	.3656	.3622
.3774	.3739	.3705	.3671	.3636	.3602
.3757	.3722	.3687	.3653	.3618	.3584
.3741	.3706	.3671	.3636	.3601	.3566
.3726	.369	.3655	.362	.3586	.3551
.3711	.3674	.3639	.3604	.3569	.3535
.3697	.3661	.3625	.359	.3555	.352
.3683	.3646	.361	.3575	.354	.3505
.3669	.3633	.3596	.3561	.3525	.349
.3657	.3619	.3582	.3547	.3511	.3476
.3645	.3607	.357	.3535	.3499	.3464
.3633	.3596	.3558	.3523	.3487	.3452
.3622	.3584	.3546	.3511	.3475	.344
.3612	.3573	.3535	.35	.3464	.3429
.3602	.3563	.3525	.3489	.3453	.3417
.3592	.3553	.3515	.3479	.3442	.3406
.3582	.3543	.3505	.3469	.3432	.3396
.3573	.3535	.3496	.346	.3423	.3387
.3565	.3525	.3486	.3451	.3415	.338
.3557	.3517	.3479	.3441	.3406	.337
.3551	.351	.347	.3434	.3398	.3362
.3543	.3502	.3462	.3426	.3391	.3355
.3536	.3495	.3455	.3419	.3382	.3346
.3529	.3488	.3447	.341	.3373	.3336
.3523	.3482	.344	.3403	.3367	.333
.3518	.3475	.3433	.3396	.3359	.3322
.3511	.3468	.3425	.3388	.3352	.3315

fig. 5. Sample data files.

```
10 HOME
20 INPUT "OVERALL DIMENSION OF DIOPLE IN MM :- ";DO
30 INPUT "INTERNAL DIMENSION OF FOLDED DIOPLE IN MM :- ";T :R=T/2
40 INPUT "DIAMETER OF ELEMENT :- ";DE
50 A=2*(DO-2*(DE+R))
60 C=2*22/7*(R+DE)
70 D=A+C
80 PRINT "TOTAL LENGTH OF MATERIAL REQUIRED :- ";D;" MM"
90 PRINT "START OF BENDS FROM CENTER OF ELEMENT :- ";A/4;"  MM."
100 END
```

fig. 6. Program to calculate material for a folded dipole.

in **fig. 2** to help with this task. This program enables you to enter the values for the file and then check the results before writing the file to the disc.

If your computer has graphic capabilities, it's a good idea to draw the curves on the screen from the saved data files. The curve should be smooth, — no bumps. *Any points out of line means you have a wrong value in the file that must be corrected.*

folded dipole construction

If you're going to use a folded dipole as the driven element, the program shown in **fig. 3**, which gives the length of tubing required and the points at which the bends forming the folded dipole should be started, will save some time. You simply enter the overall dimension of the dipole, the inside dimension of the folded dipole, and the diameter of the tubing.

The computer then specifies tubing length and the distance from the center of the element to the point to start the bend. (Once you start building the actual antenna, be sure your tube bender doesn't flatten the tube as it bends. If it does, don't forget to make allowance for this.)

when all else fails, read the screen

Several Amateurs here in VK4 have worked on the program, and we believe that it should, by now, be just about bug-free. After using the program a few times, you'll find designing Yagis much easier.

I am continually amazed at how earlier versions of the program have made their way around the world. Should you happen to have a copy of one of the earlier versions, it's important to replace it with this updated version, since the earlier versions had some errors and included none of the latest improvements.

acknowledgement

In the original articles six tubing sizes were shown on the graph; Allan, VK4KAZ, spent many hours drawing the other curves and reducing them to the values shown in the data files.

references

1. Günter Hoch, DL6WU, "More Gain with Yagi Antennas," *VHF Communications*, Volume 9 (3/1977), (4/1977); Volume 14 (3/1982).

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

the greatest contest ever staged

One of the high points of Amateur Radio is the annual ARRL DX contest, one of the oldest contests in the game. While the rules have changed from year to year, the aim of the contest is still the same: to work as much DX as possible in a given span of time.

In May, 1934, *QST* announced "the greatest DX contest ever staged!" The results of the 1933 contest had finally been tabulated and the scores were sky-high. NY1AB (Canal Zone), for example, amassed a breathtaking 25,000 points, working as many as 16 stations in an hour. And W3ZI topped the US entries with his grand total of 33,000 points. (Today it's not unusual for a contestant to pile up over a *million* points and work stations at over 250 contacts in a single hour).

What equipment were hams using in those long-gone days? The 1933 Sweepstakes provides a clue. Over 64 percent of the hams who submitted sweepstakes scores were running less than 50 watts and over 85 percent of the entries ran less than 400 watts. An amazing 15 percent of the contestants ran less than 20 watts.

In passing, *QST* noted that many high-speed CW operators were running at 25 to 30 WPM in the contest and that, in general, code speed on the DX bands was gradually increasing, year by year. Of course, there were only about 10 percent as many hams licensed in 1933-34 as there are today!

Finally it should be noted that during the 1930's, most ham gear — receiving as well as transmitting equipment — was home-made. Very few items of commercial manufacture were available, and the money to buy it was absent, for these were the years of the Great Depression.

Now I see the wheel has turned full circle. The Canadian Department of Commerce is proposing that it will require Canadian Amateurs to possess a special, advanced-type license if they want to put home-made equipment on the air!

Doesn't that seem to be placing a roadblock in the path of experimenters who want to build equipment? Experimenters should be encouraged, not harassed with the problem of getting a special license to do what should come naturally!

is your line voltage really 117 volts ac?

Have you ever checked your line voltage with a good RMS responding meter of known accuracy? You may be surprised if you do. In my case, the voltage varies over a small range from minute to minute and takes interesting swoops and dives during the day. Most of the time it runs about 123 volts, but it has dropped as low as 115 volts. Using a borrowed memory 'scope, it was found that short duration "spikes" of over 1000 volts could be observed. These were probably due to the collapsing electric field of inductive devices on the line at various points. The oil burner motor in my house, in particular, puts a nasty high-voltage spike on the power line.

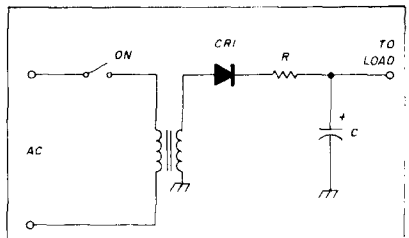


fig. 1. Resistance of transformer windings plus resistor R in series with rectifier limits inrush current to filter capacitor C.

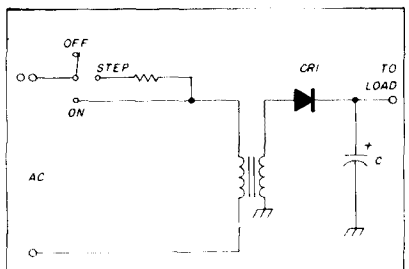


fig. 2. Shorting-type primary switch provides low voltage to supply until capacitor is charged.

Ham gear and computers can be protected from most primary line transients by virtue of inexpensive, easily available surge suppressors. These devices will protect our equipment from low-energy power line "spikes," which are the most common. More robust, industrial surge suppressors are required if you're served by a power line that also serves industrial users.

But what about transients and current surges generated within your own equipment? A surge suppressor on the power line won't help in such cases.

surge protection

It's important to incorporate surge protection in a transmitter power supply to make sure that components are not destructively overloaded during the operating cycle. Diode rectifiers and transmitting tubes are particularly vulnerable in this respect.

In the case of the diode rectifier, each time the power supply is turned on, the rectifier "sees" a low resistance short until the filter capacitor is nearly charged (**fig. 1**). The surge current through the diode to the capacitor can be several hundred amperes for a fraction of a second. In some designs a resistor in series with the rectifier is used to limit diode inrush current. The resistor, however, tends to degrade power supply voltage regulation since the operating current must flow through the resistor.

A more effective means of limiting power supply inrush current is to employ a step-start circuit (**fig. 2**), which applies low primary voltage to the supply until the filter capacitor is charged. This delay time is, typically, about one second in most cases. Once the capacitor is charged, full primary voltage can then be applied.

Various forms of inrush-limiting circuits are shown in **fig. 3**. Circuit *A* employs a variable autotransformer. The operator turns the transformer control and gradually advances the primary voltage as desired.

Circuit *B* employs a series-connected voltage dropping resistor (R) in the primary circuit, which is shorted out by a time-delay relay. There are various forms of time-delay circuits that should be of interest to the equipment builder.

time delay relay

A simple delay circuit is shown in **fig. 4A**. A 120-volt AC relay is connected so that it shorts out the series dropping resistor, R. The initial inrush current causes a voltage drop across the resistor and the relay will not close until the inrush current has decreased to a nominal value and the voltage across the relay coil is close to normal. This circuit is quite effective, but the AC relay tends to "chatter" during the delay period.

The circuit in **fig. 4B** employs a

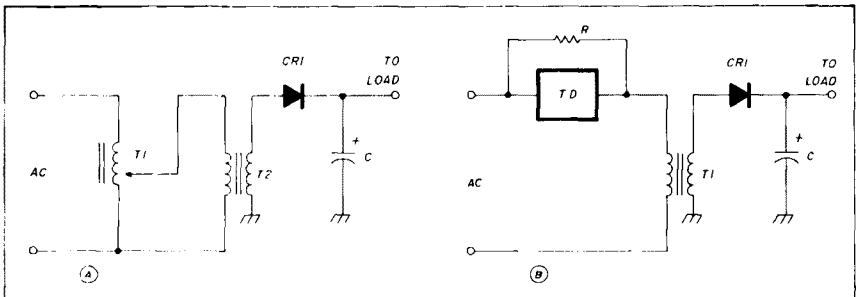


fig. 3(A). Variable autotransformer T1 permits operator to advance primary voltage slowly, thus reducing capacitor inrush current; (B) series-connected time delay relay shorts out surge resistor R after a few seconds.

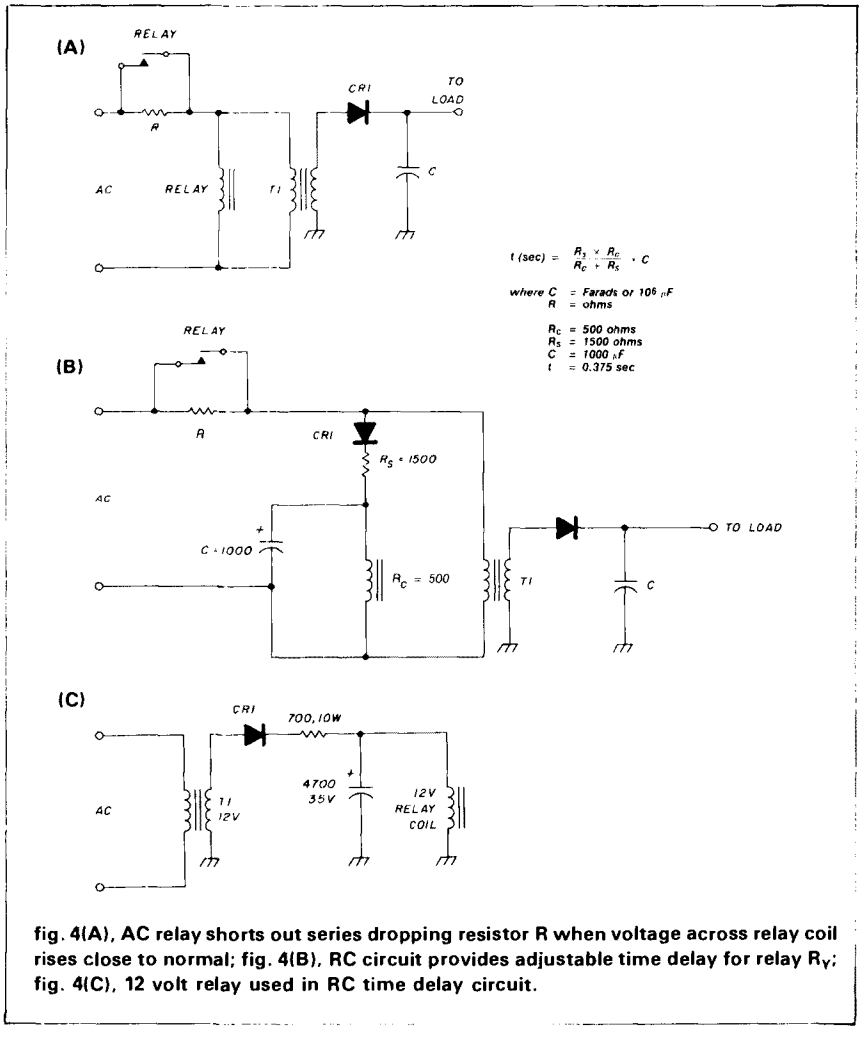


fig. 4(A), AC relay shorts out series dropping resistor R when voltage across relay coil rises close to normal; fig. 4(B), RC circuit provides adjustable time delay for relay R_Y; fig. 4(C), 12 volt relay used in RC time delay circuit.

24-volt DC operated relay with an RC time delay circuit. The line voltage is rectified and applied to the relay through a series resistor and a shunt capacitor. Besides acting as part of the time delay circuit, the capacitor also provides filtering and smoothing for the rectified AC provided by the diode.

The formula for the delay period is given in **fig. 4.** As an example, assume that the DC resistance of the relay coil (R_c) is 500 ohms and the series resistor (R_s) is 1500 ohms. The capacitor (C) has a value of 1000μF. The time delay, then, is about 0.375 seconds. The voltage across the relay coil, determined by Ohm's Law, is about 29 volts, well within the voltage tolerance of the coil.

The value of the coil voltage, series resistor and shunt capacitor can be "juggled" to provide any reasonable value of time delay. In most instances, 12 or 24 volt DC coils are used in Amateur work. The 24 volt relays are more attractive since many varieties of this type can be picked up as military surplus for a fraction of their original cost.

If desired, a 12 volt DC relay may be used, as shown in **fig. 4C.** The time delay for this circuit is about a half-second. It may be increased by increasing the size of the capacitor.

The thermostatic time delay relay consists of relay contacts mounted on a bimetallic strip which is actuated by a heater (**fig. 5**). The time delay is a function of temperature, which is controlled by the heater element. The Amperite thermostatic relay product line provides fixed time delays ranging from 2 to 180 seconds. Although the delay period can be increased by placing a resistor in series with the heating coil circuit, delays greater than 180 seconds cannot be produced. The relays resemble a receiving tube and come in 6, octal- and miniature-base designs. Although nominally 117 volts, the relays can be operated on 234 volts by the addition of a resistor in series with the heating coil.

Another form of time delay relay consists of a DC relay controlled by a small printed circuit timer built into the relay case. The Potter and Brumfield

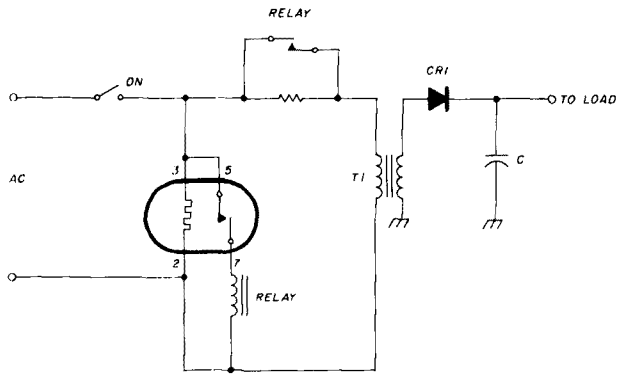


fig. 5. "Amperite" time delay relay consists of contacts mounted on bimetallic strip.

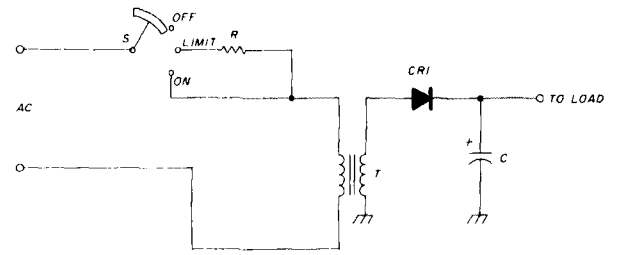


fig. 6. Progressively shorting switch provides step-start function.

type CU relay is an example of this technique. Members of the CU family of relays have an adjustable time delay period.

Compared to the thermostatic relay, the solid-state controlled relay offers the advantage of being able to be recycled immediately, while the former requires a short time interval for the thermostatic strip to cool and release the closed contacts. On the other hand, the thermostatic relay will remain closed during a short power outage, whereas the solid-state controlled relay will quickly drop open. Each relay type has its own special advantages and disadvantages, depending upon circuit requirements.

Of course, the easiest way to incorporate a surge-limiting circuit is merely to place a single-pole switch across the limiting resistor and forget about relay circuits. A progressively shorting switch (**fig. 6**) in which the positions are "off," "limit," and "on," can also be used. You can control your own time delay with these simple circuits.

filament inrush current

The time delay circuit can play an important role with regard to transmitting tubes. The tungsten filament, or heater, of a power tube has an inverse relationship between operating temperature and resistance. That is, the "cold" resistance is about one-tenth the value of the "hot" resistance. Thus, when the power tube is turned on, the filament inrush current can be as high as ten times the normal filament current.

In very large power tubes (500 kW, for example), it's often necessary to bring up the filament voltage with a motor-controlled variable voltage transformer over a period of minutes to prevent distortion of the filament due to very heavy inrush current. In the case of the lower power tubes used in Amateur service, the problem

is not as severe and the restriction of inrush current is less complex.

In many Amateur amplifiers, the filament inrush current is limited by carefully controlling the size and capacity of the filament transformer. In other words, the regulation of the filament transformer is such that the filament voltage "sags" under heavy filament current inrush conditions. This is easily accomplished in some cases by designing the filament transformer so that it's just big enough to do the job, but doesn't have extra power capacity over the amount demanded by the tubes.

In any case, filament inrush current can be limited by a time delay circuit that retards application of full filament voltage for about 3 seconds. This provides enough time for the filament to warm up and increase in resistance. Any of the delay circuits discussed previously will do the job. The series resistor value is adjusted so that about 30 percent of the rated filament voltage is applied to the tube during the delay period.

cathode warmup time

Indirectly heated cathode-type power tubes (such as the 8877 and the 4CX1000A) require a cathode warmup period before the tube is placed in operation. The warmup time required is specified by the manufacturer and depends upon cathode wattage and the physical mass of the cathode structure. The purpose of the warmup period is to ensure that the total cathode area has reached operating temperature and that there is no temperature differential across the structure. Failure to observe the warmup period can result in damage or destruction of the cathode emitting surface. Many Amateurs resent the "intolerable" time required for cathode warmup and some are tempted to cheat and start operation before the required warmup time has passed. Don't do it! You can't fool Mother Nature, so let the cathode structure reach proper operating temperature before you start calling that exotic DX station.

144 MHz EME directory

The 2-meter "moonbounce" directory of Amateurs participating in this exciting communication mode has been reprinted. For your copy, please send five first-class postage stamps or five IRCs to me at: Varian EIMAC, 301 Industrial Way, San Carlos, California 94070.

ham radio

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active antenna preamplifiers

Noiseless feedback and filter techniques improve strong signal handling capability

A recent review of commercial active receiving antennas confirmed my observation that virtually all wideband active antennas are prone to overload and intermodulation distortion in the presence of strong signals. Of eleven different consumer-grade antennas reviewed, all had the same problem.

It's difficult to imagine a high impedance input circuit with unity voltage gain that won't show some distortion when signal levels of 30 volts per meter are experienced from the local AM broadcast band, Amateur, or neighboring CB transmitters, particularly when there is no input filtering prior to amplification for the wideband case. Some improvement over present circuits can be obtained, however, by using active feedback traps for selected interference frequencies and higher power linear amplifiers for the active element. One goal is to achieve good performance from 10 kHz to 200 kHz where many of the present commercial active antennas fail to perform very well. The VLF region 10-14 kHz covering the worldwide long range Omega and Alpha navigation systems, 60 kHz WWVB and GBR time signals, and the 100 kHz LORAN-C navigation system are of interest to many Amateurs and longwave radio observers. These signals can provide very stable frequency standard references, solar activity indicators, and long range propagation data. The recent increased activity in the 160 to 190 kHz, or 1750 meter band, where 1 watt input to a 50-foot (15 meter) high antenna is allowed without license, and the use of this region for emergency government communications are of interest.

MOSFETs

Some practical details of small MOS-power FETs have been investigated with a view to application as sensitive E-field antenna preamplifiers. These are particularly useful at the VLF-LF region, where a short 1-meter whip can be made to perform as well as a much longer wire antenna. A short antenna such as a 1-meter length whip can be considered as a voltage source with a high internal impedance when coupled to a preamp input terminal. The effective source Z is equal to X_{CA}, where C_A is the antenna whip capacitance. A 1-meter whip will typically have a capacitance of 12 pF. This implies a very high input impedance preamplifier is needed, particularly at VLF-LF, where the concept is most useful. These systems are most often operated as impedance converters or voltage followers with nearly unity voltage gain. They have very high power gain in converting a signal at the high-Z source to a similar amplitude, only now at a 50- to 75-ohm receiver load.

JFET preamplifiers are most often found in consumer-grade active antenna systems where the input impedance is very high and especially where the input capacitance of the preamp system in parallel with the antenna is intended to be quite low. Capacitance at the input antenna mount and circuit will reduce the overall system gain by the resulting voltage division between the antenna and the fixed circuit input capacitance C_{IN}. **Figure 1** shows a test circuit that illustrates a way of evaluating the performance of a typical preamplifier using these parameters.

One of the common problems with all wideband active antenna systems is that of intermodulation distortion (IMD) caused by the inherent non-linearity of the preamplifier and the fact that little input filtering can be applied and still achieve a very wide bandwidth. Another problem is JFET burnout due to the relatively fragile character of the input gate structure. With these

By R.W. Burhans, 161 Grosvenor Street, Athens, Ohio 45701.

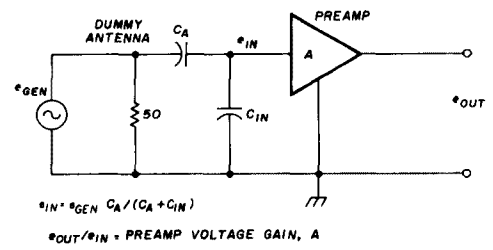


fig. 1. Test circuit.

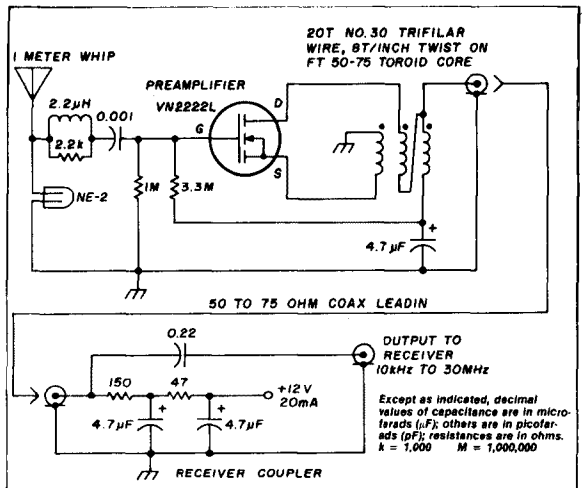


fig. 2. VLF-HF active antenna system.

facts in mind, MOS power FETs have been investigated as possible antenna preamplifiers. At first glance, the much higher input capacitance might appear to be a disadvantage, but more linear operating characteristics in the triode region, and much less tendency for gate punch-through or burnout, are of interest.

inductive feedback

A circuit that can reduce the input capacitance and improve the linearity for MOSFETs involves a "noiseless feedback" method. In this method, a portion of the drain signal is fed back either to the source or gate input via inductive transformer methods. Here the signal is fed back from drain to source with 180-degree phase reversal. The source winding feedback turns ratio determines the final output voltage gain. An experimental circuit is illustrated in **fig. 2**, where a trifilar wound toroid serves both as an output impedance matching transformer and feedback winding. As a practical matter, the greater the amount of feedback, the better the performance up to the point at which the overall circuit gain is reduced too much. With VN10KM or VN2222L VMOSFETs, the voltage gain is reduced to about −2 dB, with the FET input capacitance reduced to practical levels comparable to those of JFETs such as the J-310. **Table 1** illustrates the effects of changing the feedback winding turns ratio with respect to the output windings. The choice of a 1:1:1 ratio transformer gave the best performance in terms of minimum capacitance, gain compression level, and third order intercept as measured over the VLF-MF range from 10 kHz to 3 MHz. The 1:1 turns ratio part of the transformer used as the output provides a good match to the drain circuit of the VN2222L for a 50- to 75-ohm load at the receiver coupler. This circuit was designed originally for a maximum output level of 1V RMS or 20 mW at a 50-ohm level and actually achieved a performance of 50 mW or +17 dBm at the 1 dB gain compression level. **Figure 3** illustrates the IMD performance characteristics. This is as good or better than the performance of most presently avail-

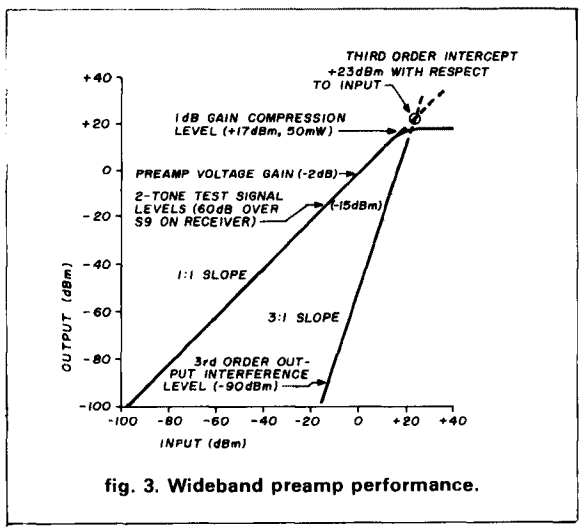


fig. 3. Wideband preamp performance.

able consumer-grade active antenna systems intended for wideband service.

Figure 4 illustrates the overall gain and phase shift for the circuit of **fig. 2**. The phase shift starting at 3 to 3.5 MHz is due to the combined effects of the length of cable connecting the coupler to the preamplifier and the output transformer resonance, where the core material tends to have less of an effect at HF. If the preamplifier is not well matched to the coupler or has excessive voltage gain, problems with spurious oscillations are sometimes noted with remote operated active antenna systems, where the total phase shift is a multiple of 2π. The very linear phase change from 10 kHz to 3 MHz for this example is useful in direction-finding applications where the signal from an E-field active antenna is combined with an H-field loop signal for a resulting cardioid directional pattern.

table 1. Feedback winding ratio effect on amplifier performance.

Feedback winding ratio (source)	Preamp voltage gain	Input capacitance (C)	E-Field attenuation (E)
0 (source ground)	+18dB	80pF	—
0.2	+10dB	70pF	—
0.4	+3dB	40pF	−16dB
1	−2dB	26pF	−18dB

sensitivity

An estimate of the antenna's sensitivity is obtained from the E-field attenuation factor noted in **table 1**. This is:

$$E_S\ (dB) = 20 \log 10 \quad \frac{e_{OUT}}{e_{IN}} \cdot \frac{1}{2} \cdot \frac{C_A}{C_A + C_{IN}} \quad (1)$$

where the factor 1/2 arises because of a given μV/m field gradient from top to bottom of a 1-meter whip. The potential developed at the preamp input terminal will be the average of the top-to-bottom field gradient. Thus the actual E-field intensity for a 1-meter whip may be estimated at the preamplifier output with a suitable calibrated receiver by applying this attenuation to the resulting signal level measured. However, such experiments are best conducted in a reasonably open area where the antenna is not obstructed by trees or nearby structures. The ultimate sensitivity of a short whip antenna such as this will depend on where it's placed with respect to the actual E-field in free space. These sensitivity numbers may appear somewhat low at the HF range compared to the antenna length, but at the VLF-LF receiving range, where large antennas are difficult or impractical to construct, this active antenna is very effective.

active notch

For severe cases of local interference due to medium wave broadcast band AM signals, an inductive feedback input trap or notch is very useful. A circuit is illustrated in **fig. 5**, which shows a tunable transformer that provides the input inductor with a feedback winding. Without feedback or with the output source winding transformer grounded, high impedance input traps like this have a very annoying peaking effect. This results from the fact that the input source impedance varies inversely with frequency, producing both resonances for these high-Z input systems. Feedback from the source winding to the series trap small coupling winding eliminates the peaking effect. Old timers may recall a related circuit with a small feedback winding used to neutralize tuned-grid, tuned-plate, single vacuum tube triode power amplifiers. The feedback

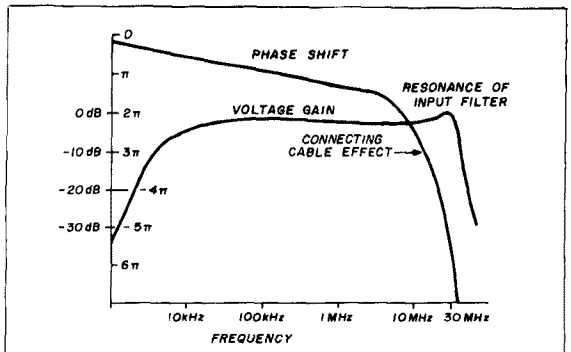


fig. 4. VN2222L antenna preamplifier.

turns ratio for the trap is very small and not critical. That is, a turns ratio of 1:10 to 1:25 will largely eliminate the peaking effect of a high-Z input trap. In choosing trap inductances, the parallel capacitor across the main series inductor (47 pF for the example of **fig. 5**) acts like an additional series capacitor at frequencies well above the notch frequency. Thus a trap of this type will always have some attenuation above the resonant notch frequency compared to the passband below the notch. For a lowpass effect, the inductor is chosen so as to be nearly self resonant and with highest possible Q, with minimum parallel capacitance. In the practical case, most all of these relatively high impedance inductors will have some distributed capacitance, which always results in some signal response above the trap frequency.

In very troublesome cases, several traps connected in series can be effective in providing a rejection filter for much of the AM broadcast band. **Figure 6** illustrates a short whip antenna system for the San Jose, California, area, where there are interference sources on frequencies 1170, 1370, 1430, 1500 and 1590 kHz. In this example the output transformer is an iron core unit normally used at audio frequencies, but which will also operate with a gradual rolloff through the 1 MHz region. The source feedback winding to the input series traps has a lower turns ratio than the examples shown in **figs. 2** and **5**, but still provides quite satisfactory gain and response for operation in the VLF-LF band.

Dual traps can also be effective for the wideband case, where the input trap is tuned to some frequency in the 1 MHz AM broadcast range and the second trap is self-resonant at a cutoff range such as 35 MHz. For this example, the output transformer should be a wideband toroid with a 0.5:1:1 or 1:1:1 turns ratio. In this case, the input trap should have a low reactance at 1 MHz, with a resonating capacitor of 100 pF or more so that the response in the passband above 1 MHz is down only −3 dB or so. In most cases with two or

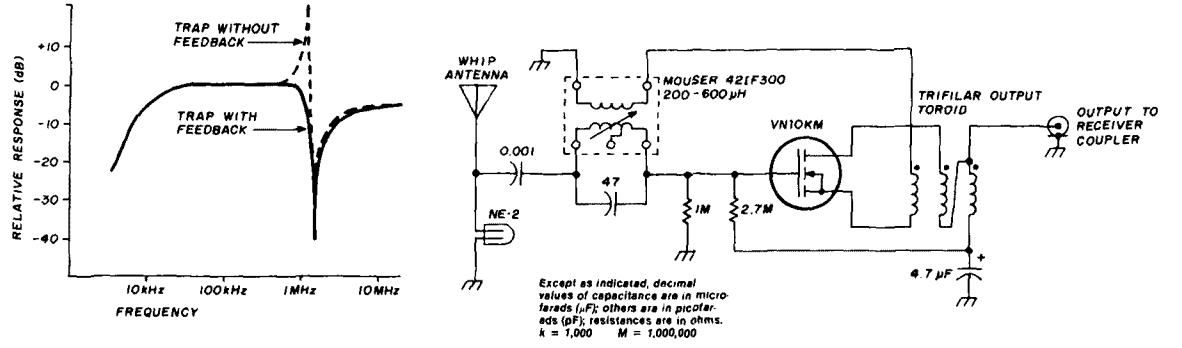


fig. 5. Noiseless feedback and input interference trap.

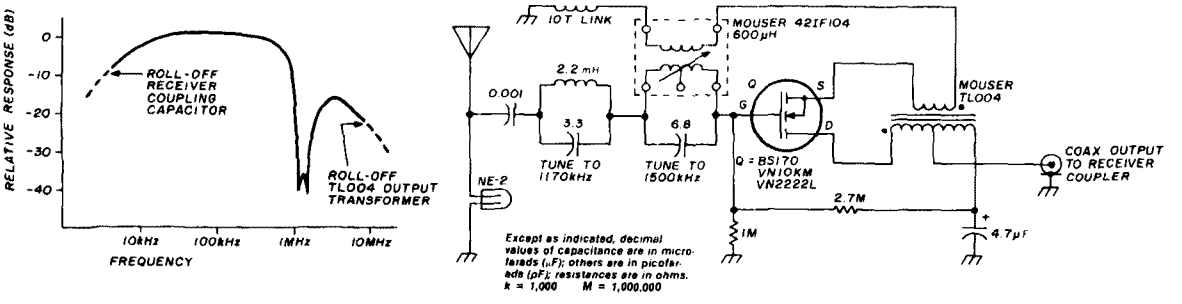


fig. 6. Power MOSFET active VLF-LF antenna with feedback traps.

more traps, the input resonator should be tuned to the lowest notch frequency with the second and succeeding traps tuned to the higher frequencies.

other applications

Another example of the utility of the noiseless feedback method is a broadband low impedance amplifier fabricated from off-the-shelf components covering the 10 kHz to 1 MHz region. The Mouser TL004 iron core transformers are used for both input and output coupling by rearranging the windings to approximate the turns ratios required for operation. A true noiseless feedback amplifier might use an alternate input transformer toroid wound as in the example of **fig. 7**. However, the TL004 iron core transformer can serve as a reasonable substitute with somewhat less power gain because of reduced turns ratios for the feedback and coupling. A proposed application of this circuit might be for a VLF-LF ferrite core loop antenna where the loop windings (N2-N3 only) substitute for the input transformer.

MOSFET selection

Several different power MOSFETs including the VN10KM, IRFDZ13, BS170, 2N7000, and VN2222L have been evaluated for these applications. One problem, however, is that they require different gate bias voltages that vary from +2 to +4 volts. The 1 megohm – 2.7 megohm voltage divider may require changing for optimum linear operation. Thus the 2N7000 series operates better with a 1 megohm – 1.5 megohm divider, producing a bias of about +3V. We have also noted that different batches of the same transistor type may have a different optimum gate bias. These power MOSFETs are designed primarily for switching service applications, but they are also generally more linear in the triode region than many zero-gate bias JFETs. The power MOSFET field has changed so rapidly over the last five years that 3rd generation types are now being introduced with still different properties. Linear circuit applications of this type are not often found in the manufacturers' literature because of the present emphasis on digital applications.

inductors

Input inductors and TL004 output transformers are available from Mouser Electronics, 1143 Woodside

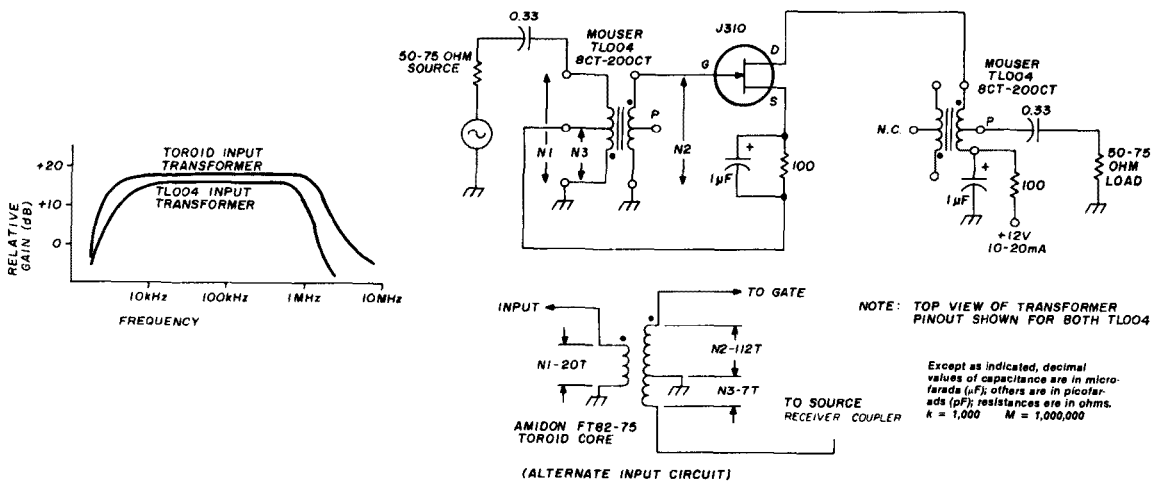


fig. 7. VHF-LF utility amplifier with noiseless feedback input method.

table 2. Series filter resonance of readily available inductors.

Mouser Part No.	Nominal Inductance (mH)	Self-Resonant Frequency (kHz)	Resonance with 10pF added (kHz)	Inductor distributed capacitance
43LJ368	68	174	134	12pF
43LH333	33	335	221	6.8pF
1120-104k	10	588	388	7.3pF
43LH268	6.8	845	516	5.2pF
1120-393k	3.9	885	614	8.2pF
1120-223k	2.2	1460	900	5.5pF
43LH215	1.5	1510	1007	7.3pF
421F104	900µH	2370	1334	5.0pF

Avenue, Santee, California 92071. The 43LHXXX series inductors are encapsulated types with values ranging from 1 mH to 33 mH. The input transformer 42IF300 is a variable cup-core type with an inductance range of about 250 to 600 µH and a turns ratio of 12:1. The 42IF104 has an inductive range of 500 to 1000 µH and a turns ratio of 22:1. These IF transformers are normally used in AM broadcast band radios but provide very satisfactory traps with the circuits illustrated. For the lower frequencies — below about 800 kHz and where a maximum lowpass effect is desired — as well as the tuned trap frequency, the LH series inductors are wound with a small coupling coil as illustrated in **fig. 8**. The polarity of the coupling is important. The circuits will oscillate with the wrong polarity of the feedback coupling of this neutralizing winding. If other inductors or transformers are substituted, a bench check for proper operation should be conducted using the input test circuit of **fig. 1**.

Table 2, lists the results of a series of measurements of these inductors used to select values suitable for traps in the AM broadcast band and part of the LF beacon band. For the best lowpass effect, the total tuning capacitance across the inductor should be kept to a minimum. The table illustrates the resonant frequencies with the coil only and with an additional 10 pF tuning capacitance. For the best lowpass effect with minimum response above the trap frequency, make the trap tuning capacitance equal to or less than the input antenna capacitance. Conversely, for the best bandpass effect, the trap tuning capacitance should be four to ten times larger than the input antenna capacitance so that the attenuation above the trap frequency is minimized.

future developments

Other experiments indicate that the feedback technique can be used with shunt as well as series inductors to eliminate the undesired peaking effect for highpass filters or traps. For future high impedance filter designs with multiple inductors, analytical methods similar to elliptic filters might be possible using these

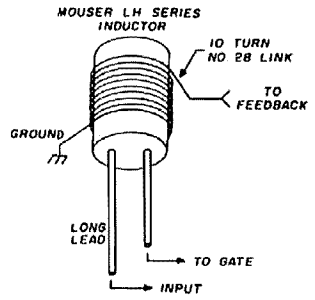


fig. 8. Winding polarity for input trap inductors.

feedback neutralizing methods of each inductor in the circuit. Perhaps a new class of high input impedance filters could be developed in which the input source impedance would vary inversely with frequency and an active feedback technique would be used to neutralize the undesired responses.

In pursuit of these ideas, the experimenter should keep in mind that all inductors have distributed capacitance which then becomes a much more critical circuit parameter. For highpass shunt filters, the active feedback polarity is reversed from the series and lowpass filters, for peaking effect reduction. Another important consideration is the filter layout where mutual coupling between two or more adjacent inductors can enhance or deteriorate a given filter performance. Close shielding of the small encapsulated type of inductors usually increases the self resonant frequency and decreases the Q because of the shorted-turns effect on the outer windings of the inductor. Combinations of feedback neutralized inductors or traps with un-neutralized resonators results in additional variety of bandpass or bandreject filters for use directly at the input from a high impedance source or antenna system.

circuit boards

For a list of experimental circuit boards and related products for these preamplifier systems, send an SASE to Burhans Electronics, 161 Grosvenor St. Athens, Ohio 45701

acknowledgements

The use of spectrum analyzer and signal generator hardware from the Avionics Enginneering Center of Ohio University in Athens, Ohio during the tests on these experimental circuits is greatly appreciated. Operational tests on several of the circuits in interference environments by Mitchell Lee, KB6FPW, have been of considerable help in improving the circuit designs.

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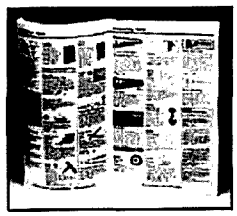
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the W2PV 80-meter quad

Design notes for a 2-element parasitic delta loop

I've been searching for a way to improve my 80-meter antenna system for several years. Along the way I acquired (from Fred Lass, K2TR) the switching system for the late Jim Lawson's two-element delta loop array. Fred also sent me a copy of Jim's engineering notes, which describe the evolution of his design over the summer of 1978. Although these notes do not define all of the antenna design parameters — particularly loop circumference and spacing — I've been able to put together a fairly complete picture of Jim's antenna system.

system description

"Quad" is a convenient misnomer; the antenna system is actually a two-element parasitic array. Each element is a nearly one-wavelength equilateral triangular loop, with one vertex (the feedpoint) at the top.

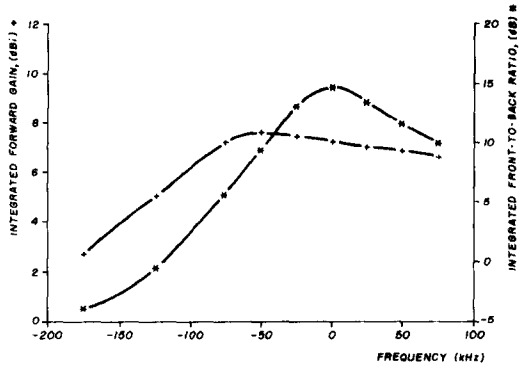


fig. 1. Performance versus relative frequency.

The loops are identical to permit convenient direction switching. The parasitic element is "tuned" to look like a reflector by shunting the feedpoint with an inductance. This arrangement provides very good gain, broad unidirectional main lobe, and moderate front-to-back (F/B) ratio, with reasonable bandwidth. Performance is not too sensitive to spacing and height, so the operating frequency can be moved from the phone band to the CW band by adding inductance in series with the loops. The array radiation resistance is quite high (near 100 ohms) so that losses in air-wound coils are negligible.

There are two principal parameters to be specified for the two-element array: the loop perimeter and the parasitic tuning inductance. W2PV selected values for these parameters through a combination of experimentation and computer analysis. His goal was to produce a resonant antenna system at the central operating frequency, with a peak in the F/B ratio versus frequency curve also at the central operating frequency. As nearly as I can infer from the notes, the loop perimeter was 254 feet (77 meters) and the reflector inductor was 4 microhenries. These parameters apply specifically to loops with apex at 152 feet (46 meters), spaced 40 feet (12 meters) apart.

gain and radiation pattern

I've modeled this antenna using the MININEC program. **Figure 1** shows the integrated gain versus frequency for fair ground. "Integrated gain" is the average value of the gain over all angles from zenith to horizon in the vertical plane containing the boom. "Integrated F/B" is the ratio of the integrated backward gain to the integrated forward gain. I believe this provides a more useful representation of antenna performance on 80 meters, where the wave angles of interest span a very large range. Note that the direc-

By Bill Myers, K1GQ, Box 501, Hollis, New Hampshire 03049

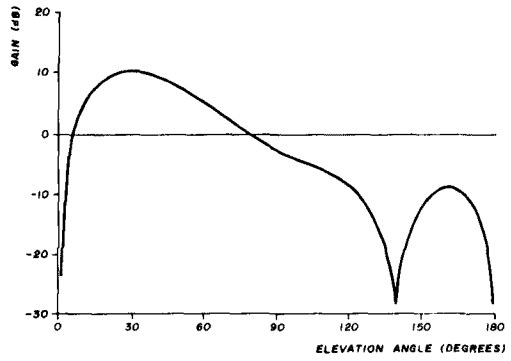


fig. 2. H-plane radiation pattern.

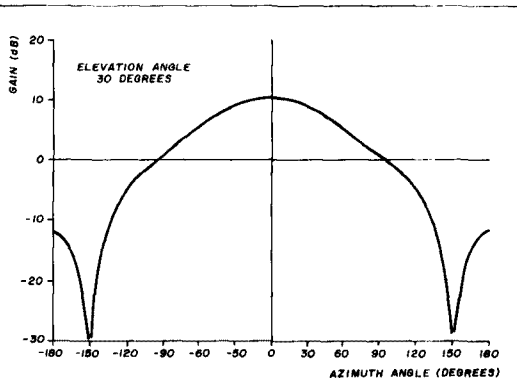


fig. 3. E-plane radiation pattern.

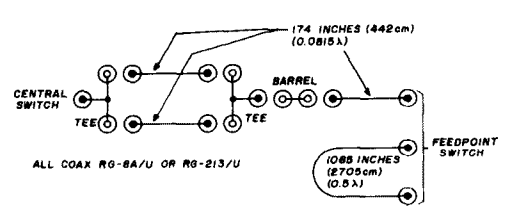


fig. 4. Coaxial matching sections.

tion of maximum gain reverses about 125 kHz below the central operating frequency, where the reflector begins to act as a director. (MININEC seems to produce reasonable radiation pattern predictions for loops, but centered at obviously wrong frequencies, so I left the center frequency undefined in **fig. 1**. Also, the MININEC predictions of input impedance are unreliable, but that's another story.)

The maximum peak gain is 11.2 dBi, 50 kHz below the center frequency. The peak gain at the center frequency is 10.8 dBi at an elevation angle of 30 degrees. These gains compare well with a rough estimate of potential gain: 3 dBi loop gain + 4 dB array gain + 6 dB ground reflection gain = 13 dBi.

Figures 2 and **3** show the H-plane and E-plane radiation patterns at the center frequency. Note that you could claim a very large F/B ratio by selecting 40 degrees and 140 degrees as the forward and backward elevation angles. The fat H-plane forward lobe provides good coverage of DX path wave angles, but less satisfactory coverage for the very high-angle (close-in) paths. The broad azimuth pattern is important because the antenna system is difficult to rotate.

matching system

The input resistance of the driven loop, at the central operating frequency, is about 100 ohms. Initially, Jim used a multi-impedance tapped balun to match this to a 50-ohm transmission line. After exhaustive tests, he concluded that this balun was not suitable and changed to the matching system shown in **fig. 4**. The half-wave balun at the feedpoint converts the 100-ohm input resistance to 25 ohms. The series-section transformer then converts 25 ohms to 50 ohms. All of these transformer sections are made of RG-8 (or RG-213) coax and are cut for 3.65 MHz. W2PV calculated the effect of the error in transformer lengths when operated at 3.5 and 3.8 MHz and decided that the consequent mismatch was unimportant. These length errors do *not* affect the array per-

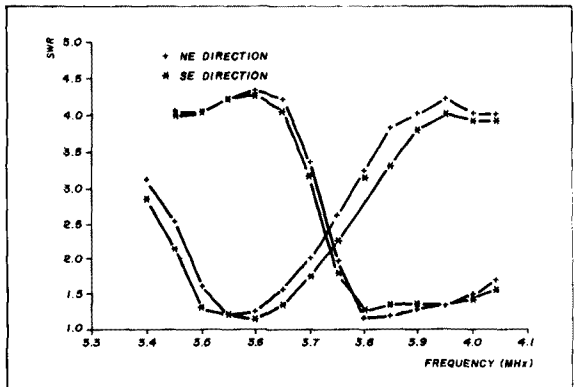


fig. 5. Measured standing wave ratio.

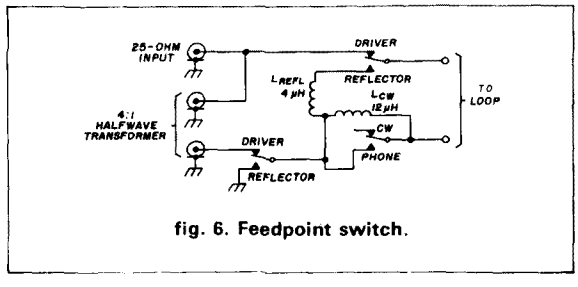
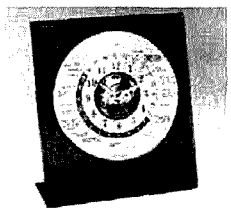


fig. 6. Feedpoint switch.

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formance, since the transformers are not part of a phasing system. The SWR data measured by W2PV is plotted in **fig. 5**; note the excellent bandwidth between the 2:1 points.

The central operating frequency is shifted down to 3.5 MHz by adding inductive reactance to both loops. Since the direction switching requires a relay box at each loop feedpoint, it's convenient to include the mode-switching relay in the same box (**fig. 6**). The complete antenna system has three relay boxes; the central box is simply a single-pole, double-throw switch that connects the main feedline to one or the other of the two loops. W2PV arranged his relay controls so that the default (i.e., no power) condition was NE/phone (with the boom running NE/SW).

closing remarks

This summer I plan to install a version of this antenna at 115 feet (35 meters) on a new tower. In addition to W2PV's parasitic arrangement, I plan to provide for feeding both loops out of phase, which yields a bidirectional pattern. The advantages of this feed are overall simplicity and small high-angle lobes. The disadvantages are somewhat lower gain and lower input resistance. If I don't knock the tower over as I cut down trees, I'll let you know how it works.

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Omni vertical collinear antennas used by FM repeaters achieve gain by vertically stacking in-phase dipoles (E-plane stack). Yagi antennas achieve gain in both E and H planes through the use of resonant parasitic elements. For Yagis with booms less than one wavelength long, almost all the gain is achieved in the H-plane. **Figure 1** shows the rate of gain improvement for both E and H plane of Yagi antennas with up to 20 dBi directive gain. As can be seen, a 3 dB improvement in the E-plane is not achieved until a total Yagi gain of approximately 16 dBi is reached.

A dipole has 2.15 dB gain over an isotropic source. All this gain is in the E-plane; the H-plane is still omni, or 360 degrees.

A three-element Yagi's −3 dB E-plane beamwidth is approximately 65 degrees, and its −3 dB H-plane beamwidth is approximately 90 degrees.

Gain improvement in the E-plane is:

$$10\ LOG\ \left\{\frac{78^\circ\ (dipole\ -3\ dB\ BW)}{65^\circ}\right\} = +0.79\ dBd$$

Gain improvement in the H-plane is:

$$10\ LOG\ \left\{\frac{360^\circ\ (omni)}{90^\circ}\right\} = +6.0\ dBd$$

Total Yagi gain is: 6.79 dBd (E and H plane gain).

If the collinear method of achieving gain in the E-plane could be combined with the Yagi's method of achieving gain in the H-plane, then the results would be rewarding. The broadside collinear "bedsprings" antenna shown in **fig. 2** does just this. The driven element is a pair of half-wave resonant dipoles (fed on the ends for high impedance) spaced a half wave apart (center to center) to achieve 2.4 dBd gain. This type of antenna achieves most of its gain in the H-plane by stacking. For those stations active on several bands and using horizontal polarization, the broadside collinear antenna takes up valuable mast space which may limit the size and number of antennas used on the other bands. A collinear radiator plus a Yagi-type parasitic antenna system seems a logical solution — thus, a collinear Yagi, or "Colagi."

Figure 3 shows the feed for a three-wide Colagi. Each of the three half-wave dipoles is in phase since the delay through the phasing network is a half-wave (180 degrees). The $0.83\lambda_0$ dipole-to-dipole center is determined by the phasing sections. This section is a coaxial line which is dielectrically loaded with polyethylene. Its 180 degree electrical length is:

$$0.659 \cdot \frac{\lambda_o}{2} \cong 0.33\lambda_o$$

physically, so the outer conductor is basically nonresonant at the design frequency.

If teflon dielectric were used, the physical length would be:

$$0.695 \cdot \frac{\lambda_o}{2} \cong 0.35\lambda_o$$

If air-spaced polyethylene were used, the physical length would be:

$$0.82 \cdot \frac{\lambda_o}{2} = 0.41\lambda_o$$

By Bob Morton, VE3BFM, P.O. Box 481, Gormley, Ontario, Canada L0H 1G0

* "Colagi" is a registered trademark of Sinclair Radio Laboratories.

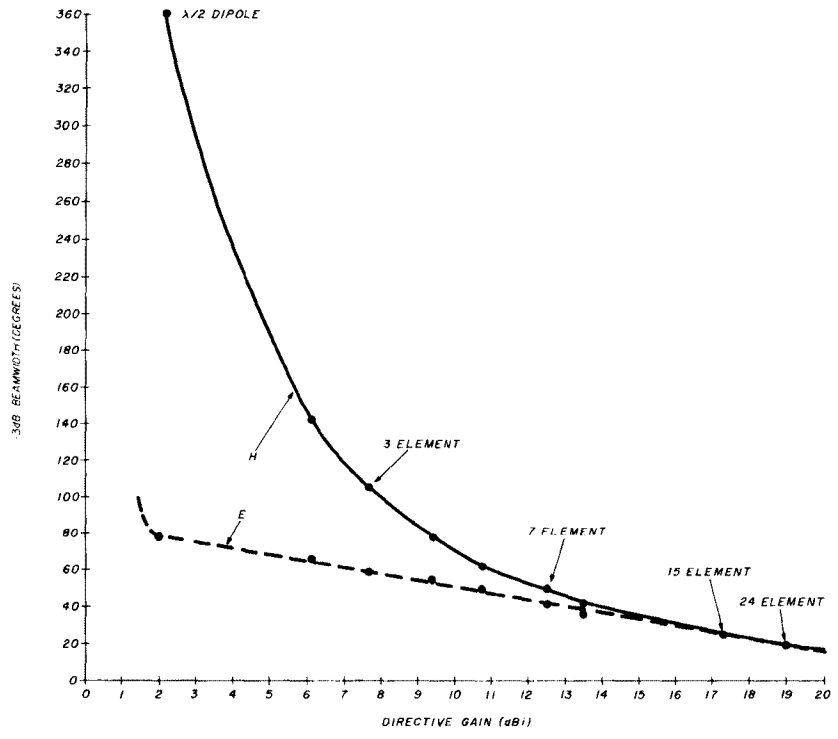


fig. 1. Dipole through 24 element Yagi –3dB beamwidth vs. directive gain.

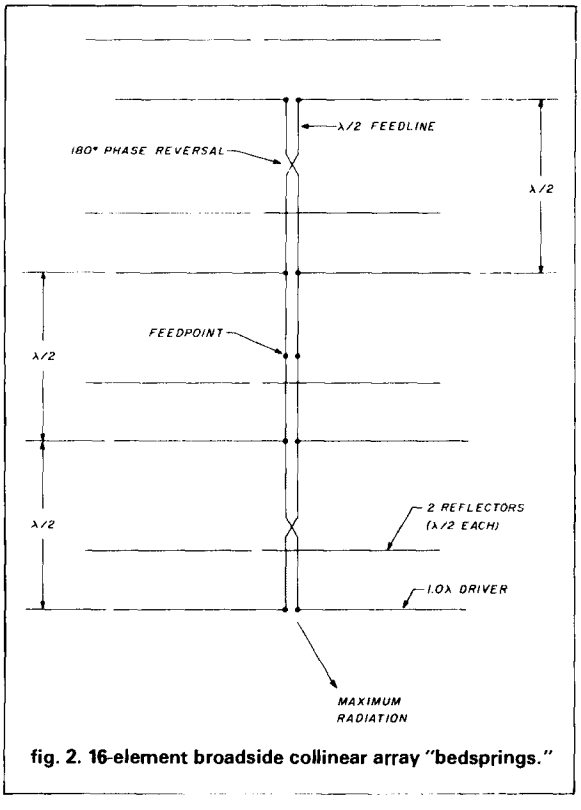


fig. 2. 16-element broadside collinear array "bedsprings."

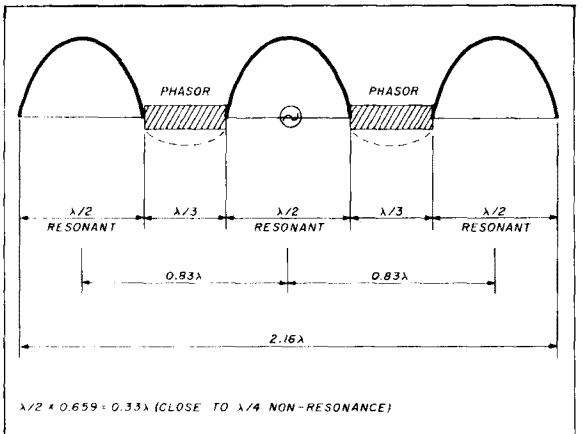


fig. 3. Current distribution for a 3-wide Colagi feed.

This length is now approaching resonance; out-of-phase radiation will occur, causing pattern distortion and reduced gain.

If the basic collinear feed (**fig. 3**) is now backed up by three separate reflectors and three equal sets of directors, a high gain antenna will be produced.

Figure 4 shows a basic seven-element Yagi using a conventional half-wave dipole feed system. E and

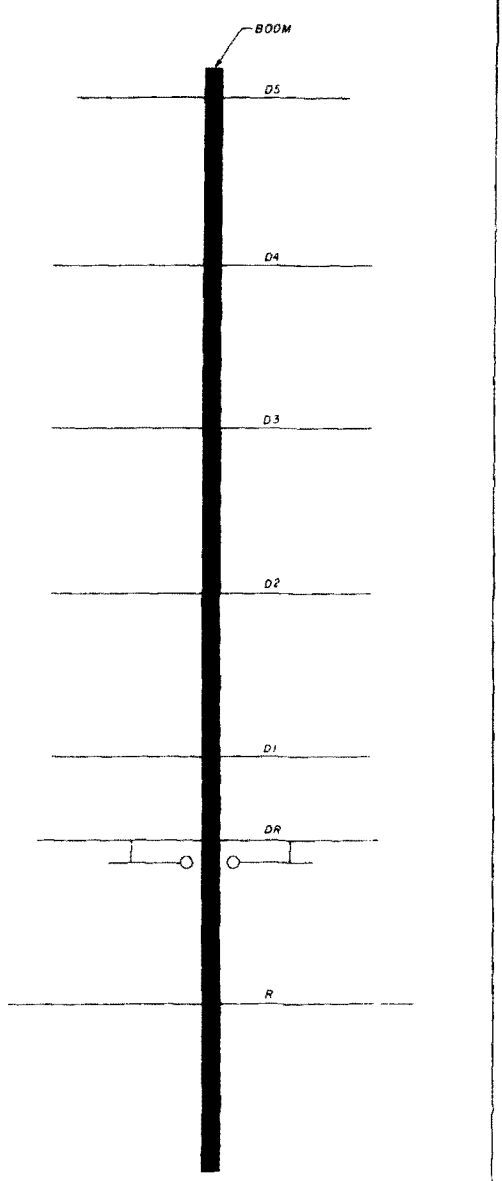


fig. 4. 7-element Yagi.

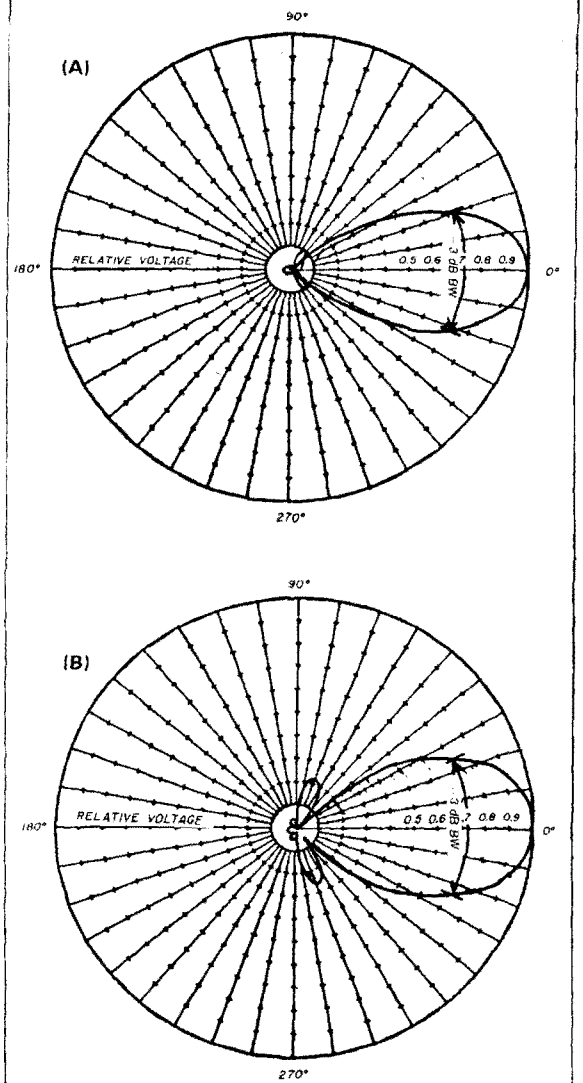


fig. 5. Seven-element Yagi −3 dB beamwidths
A) E-plane — 42 degrees.
B) H-plane — 46 degrees.

H patterns for this Yagi are shown in **fig. 5**. E-plane −3 dB beamwidth is 42 degrees and H-plane −3 dB beamwidth is 46 degrees. Gain is:

$$10\ LOG\ \left\{ \frac{32{,}000}{42 \cdot 46} \right\} - 2.15 = 10.04\ dBd$$

If this Yagi is made into a Colagi (**fig. 6**), the H-plane −3 dB beamwidth remains the same at 46 degrees. However, the E-plane −3 dB beamwidth now becomes 20 degrees. Gain now is

$$10\ LOG\ \left\{ \frac{32{,}000}{20 \cdot 46} \right\} - 2.15 = 13.26\ dBd$$

stacking separate Yagis does not yield same improvements

The same pattern could have been achieved by stacking three separate Yagis at $0.83\lambda_0$ apart in the E-plane. The disadvantage of stacking is loss in the coax and loss in the impedance matching section for the three antennas. The Colagi eliminates these particular losses. A gain improvement of 3.3 dB has been achieved and only a single feedpoint and signal boom have been used.

Figure 7 shows the array factor curve[1] for three antennas stacked at $0.83\lambda_0$. The array factor in this

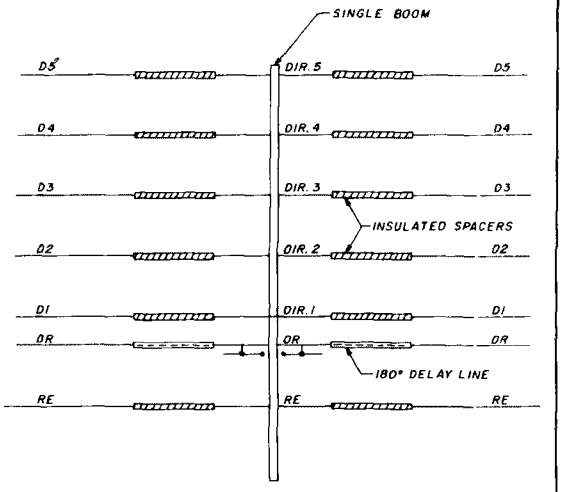


fig. 6. 3-wide Colagi.

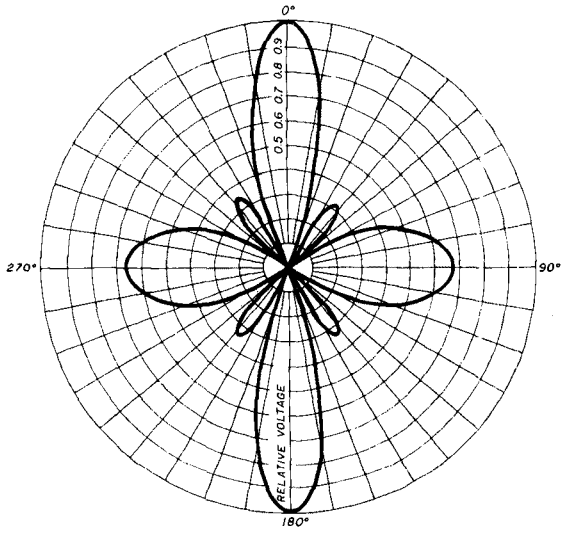


fig. 7. Array factor curve for three antennas stacked at 0.83 λ_0 spacing.

case is the resultant pattern produced by three in-line isotropic radiators of equal amplitude and phase. If this pattern is overlaid on the E-pattern of the original seven-element Yagi, the product of the two patterns will yield the resultant Colagi pattern (for the seven-element long Yagi).

The delay line phasing section is the limiting factor in determining the space between the "individual Yagis" of the Colagi.

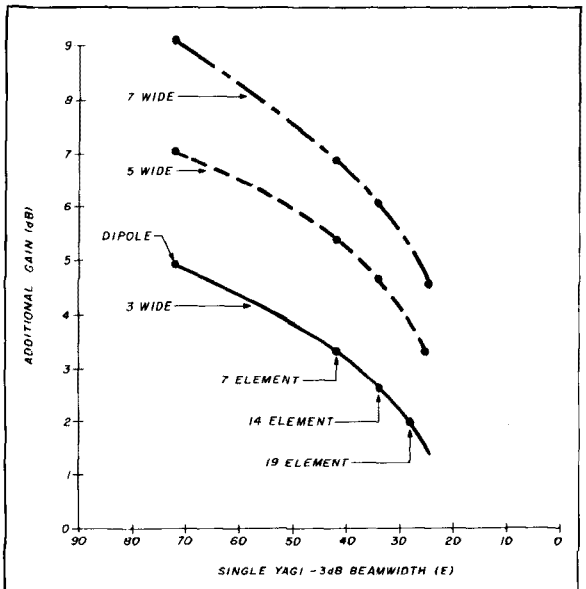


fig. 8. Gain improvement of Colagi over conventional Yagi.

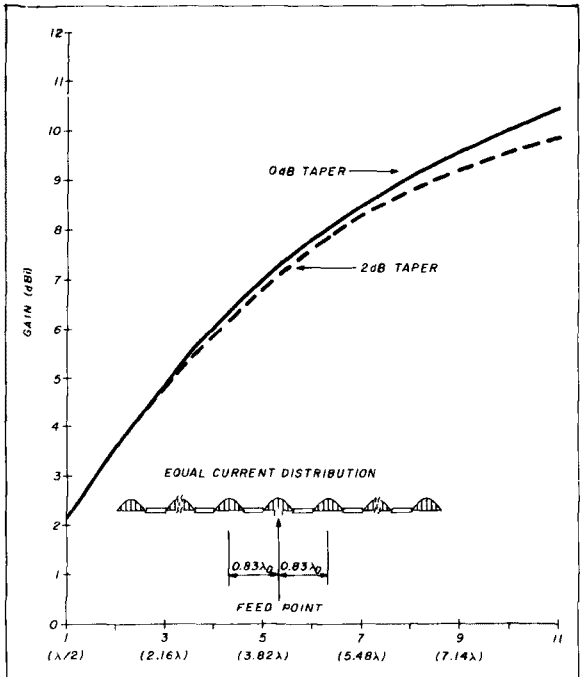


fig. 9. Gain of center fed collinear array of dipoles spaced 0.83 λ_0 with 0dB and 2dB power taper per radiating dipole.

On long-boom Yagis, the E and H patterns are approximately equal. As gain increases, the beam-widths decrease. In order to realize stacking gains, the

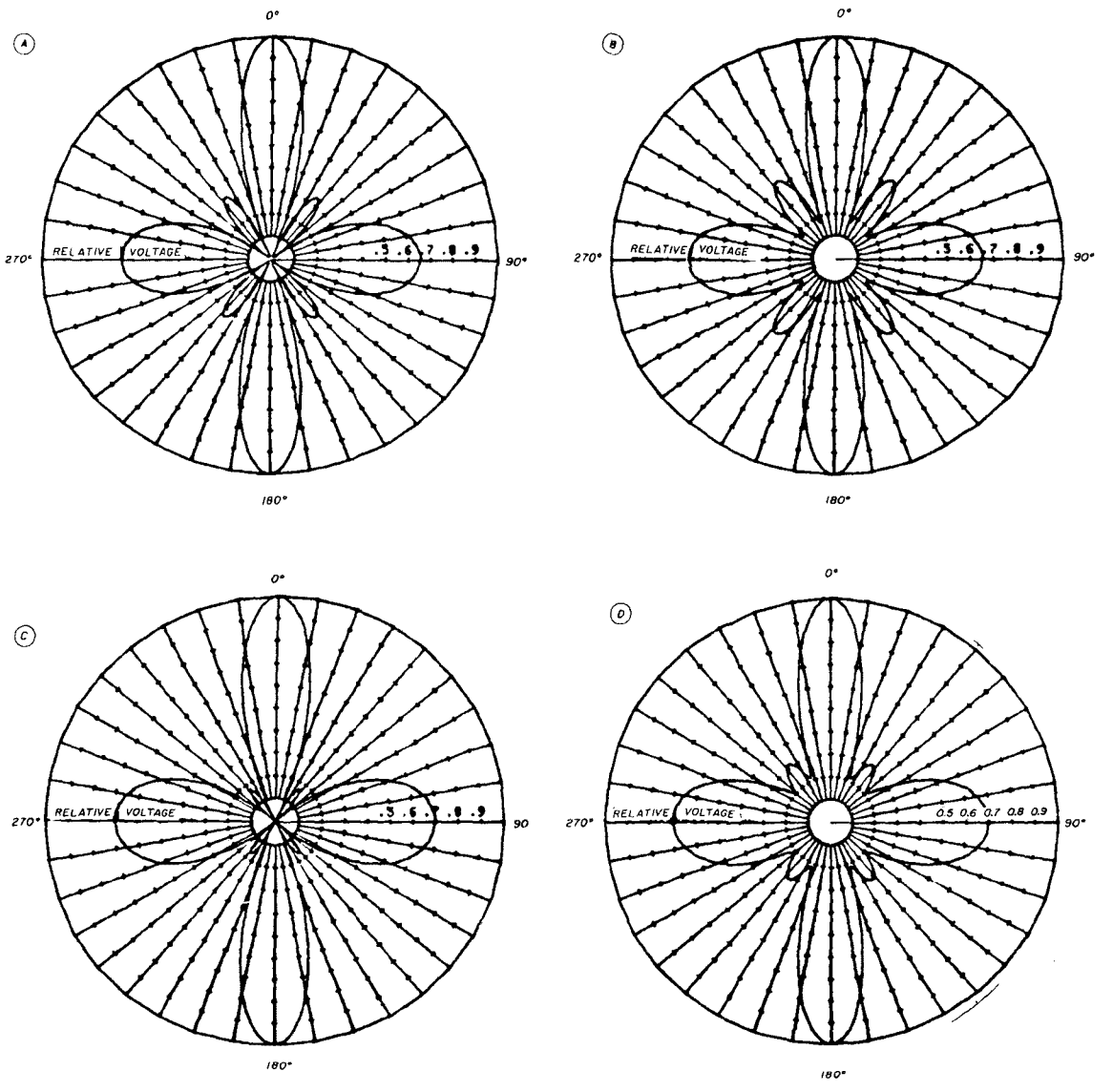


fig. 10. Phase error and power distribution affect on pattern: A) 0dB power taper and 0°, 0°, 0°, phasing. B) 0dB power taper and 30°, 0°, 30°, phasing. C) 3dB power taper and 0°, 0°, 0°, phasing. D) 3dB power taper and 30°, 0°, 30°, phasing.

spacing betwen Yagis must also increase. Using a Colagi type feed system, the spacing is fixed at $0.83\lambda_0$; therefore, an understacking effect occurs on long Yagis and all the gain available is not achieved.

The collinear feed system can be extended to five, seven, nine, . . . dipoles by increasing the number of parasitic elements accordingly.

Figure 8 shows the gain improvements to be expected for three, five, and seven-wide Colagis, based on knowing the −3 dB beamwidth in the E-plane for the single Yagi. For example, a Yagi with a 28-degree beamwidth, made into a three-wide Colagi, would see a gain improvement of only 2.0 dB, while a Yagi with a −3 dB beamwidth of 42 degrees will see a gain improvement of 3.3 dB.

unequal dipole feed reduces sidelobes

Since some power is radiated from the center dipole of a Colagi feed, there is not an equal amount left to be radiated by the other dipoles. **Figure 9** shows the gain to be expected from a center-fed collinear feed with a zero dB taper (all radiators equal in power) plus

the gain expected if a 2 dB power taper is used. On an 11-element center-fed collinear, if the center dipole power is used as a reference, then the outermost two dipoles will be 10 dB down in power. That's terrible! Right? Well, not really. The difference in gain between this condition and that with all dipole powers the same is only 0.5 dB. A slight widening of the main beam occurs when a power taper exists, but a big plus is in the reduction of the sidelobes. A very clean radiation pattern is produced. For weak signal work, having very low sidelobes is a real advantage, especially in EME work. The sky is full of noise sources and noise is additive, so why take chances with high sidelobes? A collinear fed Yagi or the Colagi is a natural for clean E-patterns.

reduced phase error pattern distortion

What about phasing errors? **Figure 10** shows the Colagi array factors:

(A) 0 dB power taper and 0°, 0°, 0°, phasing.
(B) 0 dB power taper and 30°, 0°, 30°, phasing.
(C) 3 dB power taper and 0°, 0°, 0°, phasing.
(D) 3 dB power taper and 30°, 0°, 30°, phasing.

A phasing error of 30 degrees, which would be very bad on a conventional two-Yagi stack, has virtually no effect on the resulting pattern of a Colagi with a 3 dB power taper.

By making the Yagi into a Colagi, the E-piane pattern is about 20 degrees or less for any Yagi of three elements or more. The H-plane has not been changed. Conventional stacking will approximately halve this pattern's −3 dB beamwidth each time the number of Colagis (or Yagis) is doubled. A ten-element Yagi (or Colagi) with a −3 dB beamwidth of 38 degrees in the H-plane when stacked four high will have approximately a 10 degree −3 dB beamwidth. Made into a three-wide Colagi, this array will have a gain of:

$$10\ LOG\ \left\{\frac{32{,}000}{18 \cdot 10}\right\} - 2.15 = 20.3\ dBd$$

On the 2-meter band, this gain is sufficient to hear EME echoes from a 500-watt transmitter.

A "first-level" Colagi (three-wide) on 2 meters is almost as wide as a 10-meter beam. A "first-level" Colagi on 70 cm is about 60 inches wide.

Expanding to a "second-level" Colagi (five-wide) or a "third-level" Colagi (seven-wide) continues to sharpen the E-plane pattern. For the seven-wide and up, the −3 dB beamwidth is basically the −3 dB beamwidth of the array factor. Once the Colagi is expanded to this level, the effect of adding directors does very little to the E-plane pattern. A set of reflectors behind each half-wave dipole is all that is required. However, the H-plane pattern does benefit from the addition of directors, as seen in **fig. 1**, and each builder must decide how many directors (and how long) to make the antenna.

Optimum stacking of two three-wide Colagis in the E-plane is $2.5\lambda_0$. This will give −14 dB sidelobes if a seven-element long Yagi is used as the starting point. The −3 dB beamwidth is approximately 10 degrees.

Making a seven-wide Colagi will also yield a −3dB beamwidth of approximately 10 degrees. On 432 MHz this width is 12.5 feet (3.8 meters); on 1296 MHz the width is just over 4 feet (1.22 meters). Mechanical considerations dictate just how far this type of antenna can be expanded.

At antenna measuring contests, a "figure-of-merit" is used to determine gain density. It is simply the physical dimensions of the Yagi divided into the measured gain of the antenna. My three-wide, seven-element long Colagi placed first on 1296 MHz at Dayton in 1984.

The groundwork for the Colagi is now complete. Although construction techniques will vary and final combinations will differ, all results should be rewarding.

Extra antenna gain is always desired in weak signal work. The Colagi approach might be the answer.

references

1. John D. Kraus, Ph.D., *Antennas*, McGraw Hill Book Company Incorporated, 1950, References, Appendix 17.

ham radio

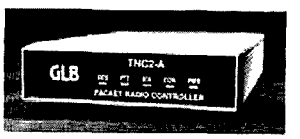
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using the antenna noise bridge

One of the most useful, inexpensive, and often overlooked test instruments is the antenna noise bridge. Over the years I've found it to be particularly useful for a variety of test and measurement applications, especially in the HF region, and those applications are not limited to the testing of antennas, which is the main job of the noise bridge. Several companies (Omega-T, Palomar Engineers, M.F.J., etc.) have produced versions of this instrument. Recently, Heath added its new Model HD-1422 (**fig. 1**) to their line-up.

Figure 2 shows a block diagram of this instrument. The bridge consists of four arms. The inductive arms (L1b and L1c) form a trifilar wound transformer over a ferrite core with L1a, so signal applied to L1a is injected into the bridge circuit. The measurement procedure consists of adjusting a 200-ohm potentiometer and a 120 pF variable capacitor. The potentiometer sets the range (from 0 to 200 ohms) of the resistive component of measured impedance, while the capacitor sets the reactive component. Capacitor C2 in the UNKNOWN arm of the bridge is used to balance the measurement capacitor. With C2 in the circuit, the bridge is balanced when C is approximately in the center of its range. This arrangement accommodates both inductive and capacitive reactances, which appear on either side of the "zero" point, i.e. the mid-range capacitance of C. When the bridge is in balance, the settings of R and C reveal the impedance across the UNKNOWN terminal (e.g. your antenna).

A reverse-biased zener diode (zeners normally operate in the reverse bias mode) produces a large amount of noise because of the avalanche process inherent in zener operation. While this noise may be a problem in many applications, in a noise bridge it is highly desirable: the richer the noise spectrum, the better. The spectrum is enhanced somewhat in the HD-1422 because of the 1 kHz squarewave modulator that chops the noise signal. An amplifier boosts the noise signal to the level needed in the bridge circuit.

The detector used in the noise bridge is an HF receiver. The preferable receiver is an AM receiver, or at least an SSB receiver with a wide IF bandwidth. Although it's quite easy to use your ears to detect the noise null that indicates bridge balance, it's best to use a receiver with an S-meter. Thus, the best receiver to use is an AM HF receiver equipped with an S-meter. If your receiver lacks an S-meter, then use an old-fashioned (analog) AC voltmeter across the receiver's speaker output. Since antennas are not always convenient to AC power, you might also consider adding "battery powered" to the list of attributes required of the receiver.

adjusting antennas

Perhaps the most common use for antenna noise bridges is finding the impedance and resonant points of an HF antenna. Connect the RECEIVER terminal of the HD-1422 to the ANTENNA input of the HF receiver through a short length of coaxial cable. The length should be as short as possible, and the characteristic impedance should match that of the antenna feedline. Next, connect the coaxial feedline from the antenna to the ANTENNA terminals on the HD-1422. You're now ready to test the antenna.

• finding impedance

Set the noise bridge resistance control to the antenna feedline impedance (usually 50 or 75 ohms for most Amateur antennas). Set the reactance control to mid-range (zero) Next, tune the receiver to the *expected* resonant frequency (f_{EXP}) of the antenna. Turn the noise bridge on and tune the receiver, looking for a noise signal of about S9 (this will vary on different receivers, and if — in the unlikely event that the antenna is resonant on the expected frequency — the S-meter reading will be much lower).

Adjust the *Resistance* control, R, on the bridge for a null — i.e., minimum noise as indicated by the S-meter. Next, adjust the *Reactance* control, C, for a null. Repeat the adjustments of the R and C controls for the deepest possible null, as indicated by the lowest noise output on the S-meter (there is some interaction between the two controls).

A perfectly resonant antenna of common Amateur designs will have a reactance reading of zero ohms and usually a resistance of 20 to 120 ohms. (There are exceptions, e.g., a resonant quarter-wave vertical has a 36.5-ohm resistive component — Ed.) Real antennas may have some reactance and a resistance that is different from 20 or 120 ohms. Impedance matching methods can be used to transform the actual resistive component to the 20 or 120 ohm characteristic impedance of the transmission line.

In general if the resistance reading is close to zero, then suspect that there's a short circuit on the transmission line and an open circuit if the

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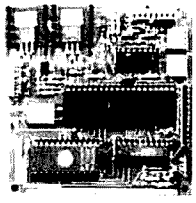
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fig. 1. Compact antenna noise bridges are useful for antenna work and component value determination.

resistance reading is close to 200 ohms. (There are exceptions — e.g., a vertical longer than $\lambda/2$ — Ed.)

For quarter-wavelength verticals and half-wavelength dipoles near resonance, a reactance reading on the X_L side of zero indicates that the antenna is too long, while a reading on the X_C side of zero indicates an antenna is too short. This convention may not hold true for other antennas, such as a vertical longer than half wavelength.

An antenna that's too long or too short should be adjusted to the correct length. To determine the correct length, we must find the actual resonant frequency, f_R. To do this, reset the *Reactance* control to zero and then *slowly* tune the receiver in the proper direction — downband for too-long and upband for too-short — until the null is found. On a high-*Q* antenna the null is easy to miss if you tune too fast. Don't be surprised if that null is out of band by quite a bit. The percentage of change is given by dividing the expected resonant frequency F_{EXP} by the actual resonant frequency (f_R), and multiply by 100:

$$Change = (f_{EXP} \times 100\%)/f_R$$

resonant frequency

Connect the antenna, noise bridge, and the receiver in the same manner as above. Set the receiver to the expected resonant frequency: i.e., 468/f for half wavelength types and 234/f for quarter wavelength types. Set the resistance control to 50 ohms or 75 ohms, as appropriate for the normal antenna impedance and the transmission line impedance. Set the reactance control to zero. Turn the bridge on and listen for the noise signal.

Slowly rock the reactance control back and forth to find on which side of zero the null appears. Once the direction of the null is determined, set the reactance control to zero and tune the receiver towards the null direction (downband if null is on X_L side and upband if on the X_C side of zero).

A less than ideal antenna will not have exactly 50 or 75 ohms impedance despite the coax impedance usually recommended, so some adjustment of R and C to find the deepest null is in order (actual values will be found throughout the noise bridge resistance range). You'll be surprised how far off some dipoles and other forms of antennas can be if they're not in "free space," — i.e., if they're close to the Earth's surface.

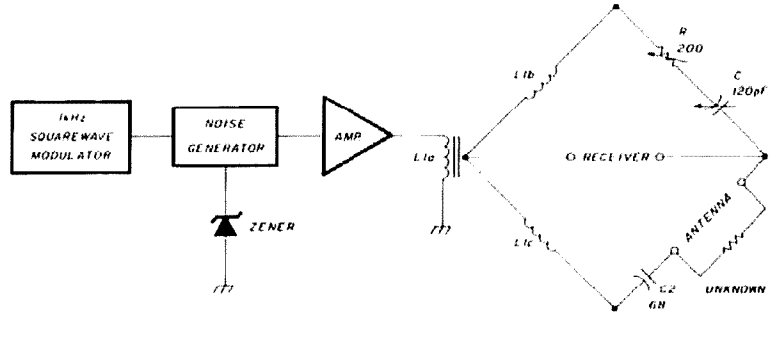


fig. 2. Antenna noise bridge circuit incorporated in the Heathkit HD-1422.

non-resonant antenna adjustment

We can operate antennas on frequencies other than their resonant frequency if we know the impedance (R and X components) and then provide a matching network to transform the impedance. Set up the receiver and noise bridge as described above and then tune the receiver to the desired operating frequency. Find the nulls for R and X (as above) and note the scale readings. The X readings are not the reactance in ohms, but rather the capacitance (0 to 60 pF). We can now calculate the normalized reactance at 1 MHz from the equations below:

$$X_C = X = \frac{159155}{68 - C} - 2340$$

or,

$$X_L = X = 2340 - \frac{159155}{68 + C}$$

Now, plug "X," calculated from one of the above, into $X_f = X/f$ where f is the desired frequency in MHz.

other applications

The Heath HD-1422 noise bridge can be used in a variety of applications. We can find the values of capacitors and inductors, determine the characteristics of series and parallel tuned resonant circuits, and calculate adjustments of transmission lines.

Some antennas and (non-noise) measurements require antenna feedlines that are either quarter wavelength or half wavelength at some specific frequency. We can use the HD-1422 to find these lengths as follows:

- Connect a short-circuit across the UNKNOWN terminals and adjust R and X for the best null at the frequency of interest (note: both will be near zero);
- Remove the short-circuit;
- Connect the length of transmission line to the UNKNOWN terminal — it should be longer than the expected length;
- For quarter wavelength lines, shorten the line until the null is very close to the desired frequency. For half wavelength lines, do the same thing, except that the line must be shorted at the far end for each trial length.

The HD-1422 can also be used to pretune an antenna tuner in order to reduce the amount of tune-up time required on the air. A previous *ham radio article* dealt with a system for doing this same job using another noise bridge.[1]

conclusion

The Heath HD-1422 noise bridge is an easily constructed, simple device that nonetheless produces useful measurement results. I recommend that all Amateurs who operate in the HF bands keep one of these instruments in their armamentareum.

reference

1. Forrest Gehrke, K2BT, "A Precision Noise Bridge," *ham radio*, March, 1983, page 50.

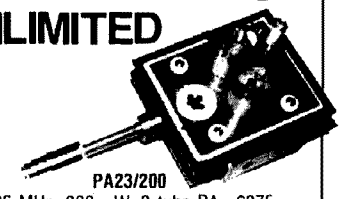
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frequency and level standard

In "A Frequency and Level Standard" (Hans Evers, PA0CX, January, page 10) the caption for fig. 3 incorrectly identifies U5 as 74SL90. The correct designation is 74LS90.

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calculations

If the impedance Z_I at the input of a transmission line of length L and characteristic impedance Z_o is known, the load impedance Z_L can be found from:

$$\frac{Z_L}{Z_o} = \frac{Z_I - jZ_o\ (tan\ B)}{Z_o - jZ_I\ (tan\ B)} \qquad (1)$$

where B = 0.367•F(MHz)•L(feet) and j is the operator $\sqrt{-1}$.

This equation can be solved for Z_I, yielding:

$$\frac{Z_I}{Z_o} = \frac{Z_L + jZ_o\ (tan\ B)}{Z_o + jZ_L\ (tan\ B)} \qquad (2)$$

Recalling the rule of complex algebra

$$\frac{a + jb}{c + jd} = \frac{ac + bd}{c^2 + d^2} + j\,\frac{bc - ad}{c^2 + d^2}\ , \qquad (3)$$

both equations can be set into forms which readily yield formulas for R_L and X_L, and R_I and X_I, respectively. Performing these transformations, one is struck by the similarity between the resulting formulas: one needs only to change the sign of the reactance, perform the calculation, and change the sign of the resulting reactance to use **eqn 1** to find the input impedance when the load impedance is known. Physically, this is because the input termination must be the conjugate of the transformed load impedance. As a practical matter, it eliminates the need to include **eqn 2** in the program.

Knowing the load impedance, the SWR can be calculated from:

$$SWR = \frac{1 + G}{1 - G} \qquad (4)$$

where G is the reflection coefficient

$$G = \sqrt{\frac{(R_L - Z_o)^2 + X_L^2}{(R_L + Z_o)^2 + X_L^2}} \qquad (5)$$

These relations may be used to calculate the SWR from the input impedance, substituting R_I for R_L and X_I for X_L, or the input SWR may be measured; however, line loss can cause the measured SWR to be significantly lower than the load SWR. The actual SWR and the total line loss can be determined from:

$$SWR\ at\ load = S_L = \frac{A + B}{A - B} \qquad (6)$$

$$SWR\ at\ input = S_I = \frac{B + C}{B - C}$$

$$Total\ loss = 10\ log\ \frac{B^2 - C^2}{B(1 - C^2)}$$

where

$$A = \frac{S_I + 1}{S_I - 1} \qquad (7)$$

$$B = 10^{M/10}$$

$$C = \frac{S_L - 1}{S_L + 1}$$

Thus, knowing the matched line loss, M, in dB, and either the input SWR or the load SWR, the total loss and the unknown SWR can be calculated.

Matched loss as a function of frequency for a variety of transmission line types can be found in numer-

By Gary E. Myers, K9CZB, 28W135 Hillview Drive, Naperville, Illinois, 60565

a typical calculation

A half-square antenna has a feedpoint impedance of 1500 + j0 ohms at 7.05 MHz. Sixty-five feet (19.81 meters) of transmission line will be required to bring it into the shack. Since only the 40-foot (12 meters) distance from the antenna to the house is in the clear, it will be difficult to use open wire all the way.

RUN the program, and enter 7.05 at the prompt. Try a *Q*-section of 450-ohm open wire, to bring the impedance down. When asked for the transmission line type, select "Other" and enter 0 for the line loss per 100 feet (30.48 meters), 450 for the line impedance, and 0.95 for the velocity factor. Enter "Y" when asked if you want a *Q*-section.

The program indicates that a *Q*-section at 7.05 MHz will be 33.15 feet (10.10 meters) in length, that the matched loss will be 0 dB, and that the electrical length is 90 degrees. It then asks for the load resistance and load reactance. After you enter 1500 and 0, respectively, it tells you that the input resistance will be 135 ohms, the input reactance will be 0 ohms, the SWR at the load will be 3.3, the input SWR will be 3.2, and that the total loss will be 0.

A *Q*-section of 75-ohm line might help at this point, so answer "Y" to the "try again?" prompt. Again enter 7.05 and choose "E" (RG11). The program says that the matched loss of this cable is 0.6 dB per 100 feet (30.48 meters). Enter 0.66 when asked for the velocity factor. Again, answer "Y" to the *Q*-section prompt, and you're told that a *Q*-section will be 23.03 feet (7.01 meters) in length with a matched loss of 0.1 dB, and that the electrical length is 90 degrees. Enter 135 and 0, respectively, when asked for the load resistance and reactance (the input impedance of the open wire section is now the load for the RG11). The program then tells you that the input resistance will be 41.6 ohms, the input reactance will be 0 ohms, the SWR at the load will be 1.8, the input SWR will be 1.7, and the total loss will be 0.1 dB.

Finally, try again, to see what the results in the shack will be if you finish up with 8.82 feet (2.68 meters) of RG213. Again enter 7.05 and choose "D" (RG8/213). The matched loss is 0.6 dB per 100 feet (30.48 meters); enter 0.66 for the velocity factor. You enter "N" at the *Q*-section prompt and are asked for the line length, which is entered as 8.82. You're told that the matched loss is 0.1 dB for 8.8 feet (2.68 meters) and that the electrical length is 34 degrees. When asked for the known quantity, respond with "B" for the load impedance. When asked for the load resistance and reactance, respond with 41.6 and 0, respectively. The computer says that the load SWR will be 1.2, the input SWR will be 1.1, the total loss will be 0, the input resistance will be 46, and the input reactance will be 8.

With just a few keystrokes, you'll have planned the feedline portion of your antenna system. You can easily do more: if you know the feedpoint impedance of the antenna at frequencies off-resonance, you can determine the 2:1 SWR points and adjust the antenna resonant frequency accordingly, to cover a preferred range of frequencies without the need for a transmatch.

ous sources. The **ARRL Antenna Book**[2] shows plots of log (frequency) vs log (attenuation) for most of the transmission lines in common Amateur use. These plots are approximately linear in the HF and VHF regions, so they can be described by $y = mx + b$ where y = *log (attenuation*, x = *log (frequency)*, m is the slope of the line, and b is the y intercept. Finding m and b for each line then allows one to calculate, with reasonable accuracy, the matched loss at any frequency up to and including VHF.

the program (fig. 1)

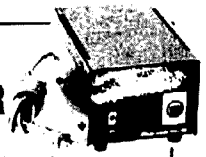
Although even the intrepid antenna experimenter cringes at the thought of going through these exercises repeatedly, it's now far less of a chore, because Transline Plus will:

- determine the input impedance, knowing that of the load;
- determine the load impedance, knowing that at the input;
- determine the apparent input SWR, knowing that at the load;
- determine the load SWR, knowing that measured at the input;
- calculate the length of a Q-section; and
- calculate the total loss.

In addition, the program automatically calculates matched line losses for several of the most common types of transmission lines in Amateur use today, and uses those values — no more thumbing through books and reading charts. It even chooses the correct Z_0.

Although the concept is straightforward, a brief

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fig. 1. "Transline Plus" program performs transmission line calculations.

```
READY.


10 REM WRITTEN BY GARY MYERS K9CZB
20 PRINT"{CLR}"
30 FORJ=1TO5:PRINT:NEXTJ
40 PRINTTAB(10)"**TRANSLINE PLUS**"
50 PRINTTAB(12)"A PROGRAM FOR"
60 PRINTTAB(4)"TRANSMISSION LINE CALCULATIONS"
70 L=LOG(10)
80 PRINT:INPUT"FREQUENCY (MHZ)";F
90 PRINT:PRINT"TRANSMISSION LINE TYPE:"
100 PRINTTAB(5)"A - RG58"
110 PRINTTAB(5)"B - RG59"
120 PRINTTAB(5)"C - 'MINI 8'"
130 PRINTTAB(5)"D - RG8/213"
140 PRINTTAB(5)"E - RG11"
150 PRINTTAB(5)"F - RG8 FOAM"
160 PRINTTAB(5)"G - OTHER"
170 GETC$:IFC$=""THEN170
180 IFC$="A"THENPOKE1558,31:LI=53.5:GOTO700
190 IFC$="B"THENPOKE1598,31:LI=73:GOTO710
200 IFC$="C"THENPOKE1642,31:LI=50:GOTO720
210 IFC$="D"THENPOKE1681,31:LI=50:GOTO730
220 IFC$="E"THENPOKE1719,31:LI=75:GOTO730
230 IFC$="F"THENPOKE1762,31:LI=50:GOTO740
240 IFC$="G"THENPOKE1799,31:GOTO260
250 GOTO170
260 PRINT:INPUT"LINE LOSS (DB/100')";LO
270 PRINT:INPUT"LINE IMPEDANCE";LI:GOTO310
280 LP=INT(LO*10)/10
290 IFINT(LO*100)/100-LP=>.05THENLP=LP+.1
300 PRINT:PRINT"  MATCHED LOSS=";LP;" DB/100'"
310 PRINT:INPUT"VELOCITY FACTOR";VF
320 PRINT:PRINT"WANT TO CALCULATE A Q-SECTION (Y/N)?"
330 GETT$:IFT$=""THEN330
340 IFT$="N"THEN370
350 QL=(246.063*VF)/F:LL=QL:QL=INT(QL*100)/100
360 PRINT:PRINT"  Q-SECTION AT";F;"MHZ=";QL;"FEET":GOTO380
370 PRINT:INPUT"LINE LENGTH (FEET)";LL
380 AL=INT(10*LL)/10
390 IFINT(100*LL)/100-AL=>.05THENAL=AL+.1
400 ML=(LO*LL)/100:MP=INT(10*ML)/10
410 IFINT(100*ML)/100-MP=>.05THENMP=MP+.1
420 PRINT:PRINT"  MATCHED LOSS=";MP;" DB FOR";AL; "FEET"
430 B=((LL/VF)*2*π)/(984.252/F):BA=B*57.3
440 BP=INT(BA):BA=INT(10*BA)/10
450 IFBA-BP=>.5THENBP=BP+1
460 PRINT:PRINT"  ELECTRICAL LENGTH=";BP;" DEGREES"
470 IFT$="N"THEN540
480 PRINT:INPUT"LOAD RESISTANCE";BR
490 PRINT:INPUT"LOAD REACTANCE";BX
500 GOSUB1270
510 PRINT:PRINT"  INPUT RESISTANCE=";RA
520 PRINT:PRINT"  INPUT REACTANCE=";XA
530 GOTO800
540 PRINT:PRINT"DESIGNATE KNOWN QUANTITY:"
550 PRINTTAB(5)"A - INPUT IMPEDANCE"
560 PRINTTAB(5)"B - LOAD IMPEDANCE"
570 PRINTTAB(5)"C - INPUT SWR"
580 PRINTTAB(5)"D - LOAD SWR"
590 GETD$:IFD$=""THEN590
600 IFD$="A"THEN750
610 IFD$="B"THEN870
620 IFD$="C"THEN1000
```

```
630 IFD$="D"THEN1080
640 GOTO590
650 PRINT:PRINT"TRY AGAIN (Y/N)?"
660 GETQ$:IFQ$=""THEN660
670 IFQ$="Y"THENPRINT"{CLR}":FORJ=1TO8:PRINT:NEXTJ:GOTO70
680 PRINT"{CLR}":FORJ=1TO10:PRINT:NEXT
690 PRINTTAB(13)"73 DE K9CZB":END
700 LA=.5376*LOG(F)/L+LOG(.42)/L:LO=10↑LA:GOTO280
710 LB=.45864*LOG(F)/L+LOG(.4)/L:LO=10↑LB:GOTO280
720 LC=.3979*LOG(F)/L+LOG(.4)/L:LO=10↑LC:GOTO280
730 LD=.515*LOG(F)/L+LOG(.21)/L:LO=10↑LD:GOTO280
740 LF=.52*LOG(F)/L+LOG(.15)/L:LO=10↑LF:GOTO280
750 PRINT:INPUT"INPUT RESISTANCE";BR
760 INPUT"INPUT REACTANCE";BX
770 GOSUB1200
780 PRINT:PRINT"  LOAD RESISTANCE=";RA;" OHMS
790 PRINT:PRINT"  LOAD REACTANCE=";XA;" OHMS
800 NP=(BR-LI)↑2+BX↑2
810 DP=(BR+LI)↑2+BX↑2
820 P=SQR(NP/DP)
830 SW=(1+P)/(1-P):SW=INT(SW*10)/10
840 PRINT:PRINT"  VSWR=";SW
850 GOSUB1120
860 GOTO650
870 PRINT:INPUT"LOAD RESISTANCE";RL
880 INPUT"LOAD REACTANCE";XL
890 NS=(RL-LI)↑2+XL↑2
900 DS=(RL+LI)↑2+XL↑2
910 G=SQR(NS/DS)
920 SW=(1+G)/(1-G):SW=INT(SW*100)/100
930 PRINT:PRINT"  VSWR AT LOAD=";SW
940 GOSUB1120
950 BR=RL:BX=-XL:GOSUB1200
960 RA=INT(RA*10)/10:XA=INT(-XA*10)/10
970 PRINT:PRINT"INPUT RESISTANCE=";RA
980 PRINT:PRINT"INPUT REACTANCE=";XA
990 GOTO650
1000 PRINT:INPUT"INPUT SWR";SI
1010 AS=(SI+1)/(SI-1)
1020 BS=10↑(ML/10)
1030 SW=(AS+BS)/(AS-BS):SW=INT(SW*10)/10
1040 PRINT:PRINT"  SWR AT LOAD=";SW
1050 CS=(SW-1)/(SW+1)
1060 GOSUB1170
1070 GOTO650
1080 PRINT:INPUT"LOAD SWR";SW
1090 GOSUB1120
1100 PRINT:PRINT"  VSWR AT INPUT=";SI
1110 GOTO650
1120 BS=10↑(ML/10)
1130 CS=(SW-1)/(SW+1)
1140 SI=(BS+CS)/(BS-CS):SI=INT(10*SI)/10
1150 IFD$="D"THEN1170
1160 PRINT:PRINT"  VSWR AT INPUT=";SI
1170 LT=(10/L)*LOG((BS↑2-CS↑2)/(BS*(1-CS↑2))):LT=INT(LT*10)/10
1180 PRINT:PRINT"  TOTAL LINE LOSS=";LT;" DB"
1190 RETURN
1200 IFCOS(B)=0THENB=1.57
1210 BB=TAN(B)
1220 D=(LI+(BX*BB))↑2+(BR*BB)↑2
1230 NA=BR*LI*(1+BB↑2):RA=(LI*NA)/D
1240 RA=INT(RA*10)/10
1250 NB=(BX-(LI*BB))*(LI+(BX*BB))+(BR↑2*BB):XA=(LI*NB)/D:
     XA=INT(XA*10)/10
1260 RETURN
1270 RA=(LI↑2*BR)/(BR↑2+BX↑2):RA=INT(RA*10)/10
1280 XA=-(LI↑2*BX)/(BR↑2+BX↑2):XA=INT(XA*10)/10
1290 RETURN

READY.
```


QST

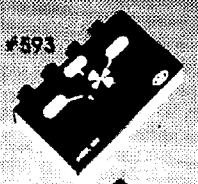
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review of the program will help to clarify things and make it easier to modify, if you wish. Since the Commodore 64 works in natural logarithms, **line 70** saves typing later on. **Lines 100-160** let you specify the type of feedline. If "other" is chosen in **line 160**, then **260** and **270** are run to let you specify the matched loss per 100 feet (30.4 meters) and the characteristic impedance; otherwise the program will do it for you. **Lines 180-240** confirm your choice with a graphic, choose the value for Z_0 and send the program to the appropriate matched loss calculation. (The loss for "mini-8" types is based on product review data for Tandy RG-8/M.[3]) **Line 250** prevents invalid entries.

Line 310 allows you to specify the velocity factor of the feedline. This could have been included in **180-240**, but since it tends to deviate from nominal values, especially in foam-dielectric types, the important part of the results may be more accurate if it is measured rather than assumed.[4] **Line 350** calculates the length of a **Q**-section if requested. **Lines 380-410** contain number-rounding operations and **420** displays matched losses. The electrical length of the line is calculated in radians and degrees, and rounded and displayed in degrees, by **lines 430-460**. If a **Q**-section is not requested, **line 470** sends the program to the next stage.

Lines 550-630 let you choose your calculation, based on the known quantity, and send the program to the appropriate routines and subroutines. **Line 1200** prevents "division by zero" errors when the line length is exactly 90 degrees.

After typing in the program, be sure to **SAVE** it before **RUNning**, in case you've made an error that will cause a system lockup. For those with little programming experience, note that the CLR in curly brackets in **lines 20, 670**, and **680** means a **SHIFTed CLR/HOME** key; this shows on the screen as a reversed video heart symbol.

results

The program is designed to be easy to use. Just follow the prompts as they appear on the screen. Prompts that ask for a single letter response are executed immediately by **GET** statements; those that ask for numbers are **INPUT** statements, and require a **RETURN** for execution. If you want a hard copy of your results, add **PRINT#** statements after each appropriate screen **PRINT**.

Some cautions are in order. The impedance transformation calculations don't account for line loss. Normally this is of little consequence in well-designed systems, but it may introduce significant errors if the total loss exceeds 2 dB or so. The exponential relations developed by KØOP[5] might be substituted for the calculations in **lines 1210-1250**, if better accuracy is desired for very lossy situations. The calculations for

matched loss are based on the assumption that the log (frequency)/log (attenuation) relationship is linear; this is only approximately true for some types, especially in the UHF region. Rounding operations and some minor approximations cause the program to yield SWR and line loss results that may be inconsistent in the first decimal place.

A typical calculation is illustrated in the sidebar.

This program has taken much of the tedium out of my antenna system studies, and makes "what if?" exercises a real joy. If you don't care to type the program, send me a formatted disk and $5, and I'll provide you with a copy.

references

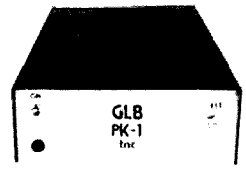
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3. Jerry Hall, K1TD, "Tandy Wire and Cable RG-8/M Coaxial Cable," *QST*, December, 1980, page 49.
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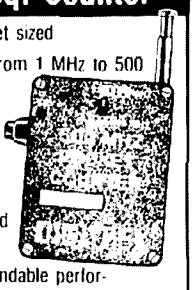
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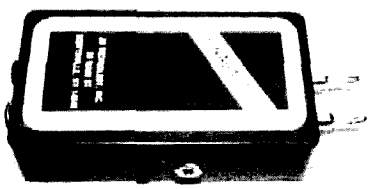


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hf ground wave propagation

A computer program for calculating expected range

While some hams are busy chasing rare DX, working "skip" via F2 ionized atmospheric layers, many others are busy sending SSTV signals across town to friends or communicating with relatives in adjacent states by using "ground wave" propagation. If you're the latter, and you'd like to discover an armchair method of determining what your expected ground wave communications range should be, then read on.

This article describes a computer program, written in BASIC for the Commodore 64 or 128PC — but easily converted to other computers — which will predict your range in miles as you experiment with transmitter power, receiver sensitivity, polarization, and antenna heights and gains.

The program is completely menu-driven for ease of operation. Two typical screens are shown in **figs. 1** and **2.** The first is used for selection of frequency band, equipment characteristics, and polarization, and the latter is used in the selection of antenna heights and gains. The typical output format is shown in **fig. 3,** and a routine is included to dump all applicable information to your printer.

propagation curves

Although propagation curves have been available for about 40 years, most hams have either not been aware of them or haven't known how to use them. The classic curves developed by Bell Labs cover 200 kHz to 600 MHz over distances of 0.5 to 1000 miles (0.8 to 1600 km) and are arranged in six sections covering propagation over poor soil, good soil, and sea water for vertical and horizontal polarizations, and include ground-to-air data to 40,000 feet (12,000 meters).[1] Typical inputs and outputs are expressed in terms of 1 kW transmitted from a grounded whip and units of field strength in dB above 1 microvolt per meter, however, one must be wise in the ways of antenna conversions to use them. Although propagation predictions at VHF and UHF are fairly straightforward because antennas at these frequencies are usually mounted many wavelengths above earth, where ground effects are negligible,[2,3] they are very complicated at HF, where antennas are usually located within a few wavelengths of ground. Under these conditions, actual antenna directivity and efficiency are a function of polarization and are directly affected by soil conductivity, and other factors.

The user-friendly menu-driven computer program described below utilizes data taken from portions of selected curves in reference 1. Ground wave propagation data are included for the 3.5, 7, 14, 28, and 50 MHz bands, with separate information for both vertical and horizontal polarizations for three different ground conductivities.

program description

The menu-driven HF ground wave propagation program includes a sufficient number of INPUT, PRINT, and REMARKS statements that should make it self-explanatory to most users. The program LISTing in BASIC is shown in **fig. 4.** The following description is for those who wish to follow the program flow line by line.

Lines 10 through **40** display the program title on the screen and provide you an opportunity to select any combination of border, screen, and letter colors desired; you're not stuck with the Commodore 64 default conditions. In **lines 50-74** you choose whether to work with receiver sensitivity and transmitter power in microvolts and watts or in dBm (decibels relative to 1 milliwatt). This is for your convenience only; the program converts either input to the other and displays both as an output.

Line 76 branches to the program data in the subroutine beginning at **line 9000** and running to the end of the program. (Details of the data format will be given later for interested programmers.) **Lines 100-115** then branch to subroutines for initial inputs of the parameters described below. The subroutine in **lines 600-624** prints a menu on the screen (see **fig. 1**) and requires you to select your operating frequency band.

By Lynn A. Gerig, WA9GFR, R.R.#1, Morgan Road, Monroeville, Indiana 46773

```
SELECT FREQUENCY BAND FROM MENU

          1 =  3.5 MHZ
          2 =  7   MHZ
          3 = 14   MHZ
          4 = 28   MHZ
          5 = 50   MHZ

WHAT IS YOUR CHOICE? 5


INPUT XMTR POWER (IN WATTS)? 100
RCVR SENSITIVITY (IN MICRO-VOLTS)? .8

VERTICAL OR HORIZONTAL POLARIZATION
(V OR H)? H


SELECT PROPAGATION PATH FROM MENU

 1 = POOR SOIL
 2 = GOOD SOIL
 3 = SEA WATER

WHAT IS YOUR CHOICE? 2
```

fig. 1. Menus for selection of frequency band, radio parameters, polarization, and path type.

```
CHOOSE ANTENNA FEEDPOINT HEIGHT ABOVE
GROUND FROM THE FOLLOWING MENU:


1 =  10'    4 =  40'    7 =  80'
2 =  20'    5 =  50'    8 = 100'
3 =  30'    6 =  60'    9 = 150'


SELECT HEIGHT OF TRANSMIT ANTENNA? 3
SELECT HEIGHT OF RECEIVE ANTENNA? 5


ENTER GAIN OF XMIT ANTENNA (IN DB)? 12
ENTER GAIN OF RCV ANTENNA (IN DB)? 7.5


ENTER TOTAL LOSSES AT TRANSMITTING AND
RECEIVING ENDS OF THE LINK.  INCLUDE
COAX CABLE LOSSES, ETC. (IN DB)?  4
```

fig. 2. Menus for selection of antenna parameters.

The subroutine in lines **700-740** asks for receiver sensitivity and transmitter power output. You'll either be prompted for an input in dBm or in microvolts and watts, depending upon your preference selected in lines **70-74**. Your input is converted to both units, which will be displayed later.

The subroutine in lines **800-860** displays menus on the screen requiring you to select polarization (vertical or horizontal) and type of propagation path (poor soil, good soil, or sea water), also shown in **fig. 1**.

All antenna parameters are input from the subroutine in lines **900-995**. You must first select the heights of both the receiving and transmitting antennas from the menu (see **fig. 2**). The program includes data for antenna elevations above ground from 10 to 150 feet (3.04 to 45.7 meters) in nine discrete increments. In addition, if you select vertical polarization, an additional option for specifying your antenna at ground level is included. Next you're asked to input antenna gains in dB. Finally, you must input system losses. These would include coaxial cable losses, antenna matching network losses, and antenna losses due to a poor ground radial system, unless you took these into consideration when entering radio set parameters. For example, if you have a 100-watt transmitter and have 6 dB of coax cable and antenna losses, either use 100 watts as your power level and include 6 dB in losses (plus losses at the receiving end, of course), or use 25 watts (actual radiated power) with 0 dB losses.

The main program calculations and outputs occur in lines **200-335**. First, path loss vs distance and antenna height gain data for your operating conditions are selected based upon your inputs of frequency, polarization, soil type, and antenna heights. An equivalent "path" in dB is calculated in line **240**. This number is basically the difference in dB between the transmitter power output and receiver sensitivity, with corrections for antenna gains and system losses, plus an equivalent antenna "height gain" for antennas at other than "reference height." Data stored in the pro-

```
GROUND WAVE PROPAGATION AT 50 MHZ
HORIZONTAL POLARIZATION OVER GOOD SOIL

TRANSMITTER POWER OUT: 50      DBM
                       100     WATTS

RECEIVER SENSITIVITY: -108.9   DBM
                       .8      UV

TRANSMITTING ANTENNA:  12 DB GAIN
                       AT 30 FEET

RECEIVING ANTENNA:     7.5 DB GAIN
                       AT 50 FEET

SYSTEM LOSSES:         4 DB

MAXIMUM EXPECTED RANGE:  66.6 MILES

----------------------------------------
F=NEW FREQ BAND          R=RUN AGAIN
A=MODIFY ANTENNAS        P=PRINTER DUMP
X=MODIFY R/T SENS/PWR    Q=QUIT
N=NEW PATH OR POLARIZATION
```

fig. 3. Typical program output format.

fig. 4. Propagation program list in BASIC for the Commodore 64.

```
2 CL$=CHR$(147):D$="N":LT$=CHR$(157):DN$=CHR$(17)
10 PRINTCL$DN$DN$"    3.5-50 MHZ GROUND-WAVE PROPAGATION
14 PRINTDN$"        PROGRAM  FOR THE COMMODORE 64
16 PRINTDN$DN$DN$DN$"                V1.0 C 1985 BY
18 PRINTDN$"          LYNN A. GERIG, WA9GFR
20 FORJ=1TO6:PRINT:NEXT
22 PRINT"TO CHANGE BORDER, SCREEN, OR LETTER
24 PRINT"COLORS, PRESS B, S, OR L, RESPECTIVELY.
26 PRINTDN$"TO EXIT TO PROGRAM, PRESS <RETURN>.
30 GETA$:IFA$=""THEN30
32 IFA$="B"THENPOKE53280,PEEK(53280)+1AND15
34 IFA$="S"THENPOKE53281,PEEK(53281)+1AND15
36 IFA$="L"THENPOKE646,PEEK(646)+1AND15:GOTO10
38 IFA$=CHR$(13)THEN50
40 GOTO30
50 PRINTCL$DN$"THIS PROGRAM CALCULATES EXPECTED RANGES
52 PRINT"FOR FREQUENCY BANDS OF 3.5, 7, 14, 28,
54 PRINT"AND 50MHZ.  THE APPROXIMATE DYMANIC
56 PRINT"RANGE IS FOR PATH LOSSES OF 100 TO 200
58 PRINT"DB, COVERING MOST APPLICATIONS FOR RCVR
60 PRINT"SENS OF .2 TO 10 MICRO-VOLTS AND XMTR
62 PRINT"POWER OF .1 TO 1000 WATTS.  PROGRAM
64 PRINT"COVERS ANTENNA HEIGHTS FROM GROUND
66 PRINT"LEVEL TO 150 FEET.
70 PRINTDN$DN$"PROGRAM DEFAULTS TO RCVR SENS AND XMTR
72 PRINT"PWR IN DBM.  WOULD YOU RATHER WORK WITH
74 PRINT"MICRO-VOLTS AND WATTS (Y=YES)";:INPUT D$
76 GOSUB9000:REM READ DATA
100 GOSUB600:REM SELECT FREQUENCY BAND
105 GOSUB700:REM SELECT XMTR PWR & RCVR SENS
110 GOSUB800:REM SELECT PROPAGATION PATH
115 GOSUB900:REM SELECT ANTENNA PARAMETERS

200 REM MAIN PROGRAM CALCULATIONS
205 IF(H1=00RH2=0)ANDPO=2THENGOSUB900
210 H$=H$(PO,PP,FB):REM DATA FOR SELECTED POLARIZATION, PATH & FREQ
215 FORJ=1TO12:X=6*(J-1):P(J)=VAL(MID$(H$,X+1,3))
220 D(J)=VAL(MID$(H$,X+4,3)):NEXT
225 IFD(12)=999THEND(10)=1050:D(11)=1050:D(12)=1050
230 HX=HG(PO,PP,FB,H1):REM XMIT ANTENNA HEIGHT GAIN
235 HR=HG(PO,PP,FB,H2):REM RCVR ANTENNA HEIGHT GAIN
240 PL=PD-RD+HX+HR+GX+GR-LL:REM PATH IN DB
245 PRINTCL$"GROUND WAVE PROPAGATION AT"F(FB)"MHZ
250 PRINTPO$(PO)" POLARIZATION OVER "PP$(PP)
255 PRINTDN$"TRANSMITTER POWER OUT:"PD;TAB(30);"DBM
260 PRINTTAB(22)PW;TAB(30)"WATTS"
265 PRINTDN$"RECEIVER SENSITIVITY:  "RD;TAB(30)"DBM
270 PRINTTAB(22)RMTAB(30)"UV
275 PRINTDN$"TRANSMITTING ANTENNA:  "GX"DB GAIN
280 PRINTTAB(23)"AT"H(H1)"FEET
285 PRINTDN$"RECEIVING ANTENNA:     "GR"DB GAIN
290 PRINTTAB(23)"AT"H(H2)"FEET
295 PRINTDN$"SYSTEM LOSSES:         "LL"DB"DN$
300 IFPL<P(1)THENPRINT"RANGE NOT IN PROGRAM:   <"D(1)"MILES":GOTO400
305 IFPL>=P(12)THENPRINT"RANGE NOT IN PROGRAM:   >"D(12)"MILES":GOTO400
320 FORJ=1TO11:X=J+1
325 IFPL>=P(J)ANDPL<P(X)THENDI=D(J)+(D(X)-D(J))*(PL-P(J))/(P(X)-P(J))
330 NEXT:DI=INT(10*DI+.5)/10
335 PRINT"MAXIMUM EXPECTED RANGE:  "DI"MILES
400 PRINTDN$LL$
402 PRINT"F=NEW FREQ BAND           R=RUN AGAIN
404 PRINT"A=MODIFY ANTENNAS         P=PRINTER DUMP
406 PRINT"X=MODIFY R/T SENS/PWR     Q=QUIT
408 PRINT"N=NEW PATH OR POLARIZATION";
410 FORJ=1TO10:GETA$:NEXT
412 GETA$:IFA$=""THEN412
413 IFA$="P"THEN430
414 PRINTDN$
415 IFA$="R"THEN100
416 IFA$="Q"THENSYS65126:REM RESTORE
418 IFA$="N"THENGOSUB800:GOTO200
420 IFA$="A"THENGOSUB900:GOTO200
422 IFA$="X"THENGOSUB700:GOTO200
424 IFA$="F"THENGOSUB600:GOTO200
426 GOTO412
430 REM SCREEN-DUMP TO PRINTER
432 OPEN3,3:OPEN4,4:PRINTCHR$(19);:PRINT#4:PRINT#4,LL$
434 FORJ=0TO759:GET#3,A$:PRINT#4,A$;:NEXT:PRINT#4,LL$
436 CLOSE4:CLOSE3:FORJ=1TO5:PRINT:NEXT:GOTO412
```


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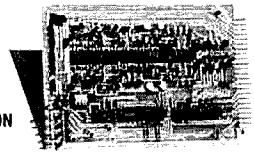
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```
600 PRINTCL$"SELECT FREQUENCY BAND FROM MENU"DN$
602 PRINTTAB(9)" 1 =  3.5 MHZ
604 PRINTTAB(9)" 2 =  7   MHZ
606 PRINTTAB(9)" 3 = 14   MHZ
608 PRINTTAB(9)" 4 = 28   MHZ
610 PRINTTAB(9)" 5 = 50   MHZ"DN$
620 INPUT"WHAT IS YOUR CHOICE";FB
622 IFFB<>1ANDFB<>2ANDFB<>3ANDFB<>4ANDFB<>5THEN600
624 PRINTDN$:RETURN
700 IFD$="Y"THENINPUT"INPUT XMTR POWER (IN WATTS)";PW:GOTO710
702 INPUT"INPUT XMTR POWER (IN DBM)";PD
704 PW=(PD-30)/10:PW=10^PW
706 IFPW>=1THENPW=INT(PW$10+.5)/10:GOTO720
708 IFPW<1THENPW=INT(PW$1000+.5)/1000:GOTO 720
710 PD=10$LOG(PW)/LOG(10)+30:PD=INT(PD$10+.5)/10
720 IFD$="Y"THENINPUT"RCVR SENSITIVITY (IN MICRO-VOLTS)";RH:GOTO730
722 INPUT"RCVR SENSITIVITY (IN DBM)";RD
723 IFRD>0THENPRINT"<1 MW IS A NEGATIVE NUMBER":GOTO722
724 RH=(RD+107)/20:RM=10^RM
726 IFRM>=1THEN RM=INT(RM$10+.5)/10:GOTO740
728 IFRM<1THEN RH=INT(RM$100+.5)/100:GOTO740
730 RD=20$LOG(RM)/LOG(10)-107:RD=INT(RD$10+.5)/10
740 RETURN
800 REM SELECT POLARIZATION & PROPAGATION PATH
805 PRINTDN$"VERTICAL OR HORIZONTAL POLARIZATION
810 INPUT"(V OR H)";PO$
815 IFPO$="V"THENPO=1:GOTO830
820 IFPO$="H"THENPO=2:GOTO830
825 GOTO810
830 PRINTDN$DN$"SELECT PROPAGATION PATH FROM MENU"DN$
835 PRINT" 1 = POOR SOIL
840 PRINT" 2 = GOOD SOIL
845 PRINT" 3 = SEA WATER"DN$
850 INPUT"WHAT IS YOUR CHOICE";PP
855 IFPP<>1ANDPP<>2ANDPP<>3THENGB30
860 RETURN
900 REM ANTENNA PARAMETERS
905 PRINTCL$"CHOOSE ANTENNA FEEDPOINT HEIGHT ABOVE
910 PRINT"GROUND FROM THE FOLLOWING MENU:":PRINTDN$
915 IFPO=1THENPRINT"O =  GROUNDED VERTICAL"DN$
920 PRINT"1 =  10'    4 =  40'    7 =  80'
925 PRINT"2 =  20'    5 =  50'    8 = 100'
930 PRINT"3 =  30'    6 =  60'    9 = 150'"DN$DN$
935 INPUT"SELECT HEIGHT OF TRANSMIT ANTENNA";H1:H1=INT(H1)
940 INPUT"SELECT HEIGHT OF RECEIVE ANTENNA";H2:H2=INT(H2)
945 IFH1<0ORH2<0ORH1>9ORH2>9THENPRINT"NOT IN MENU":GOTO935
950 IFPO=2AND(H1=0ORH2=0)THENPRINT"NOT IN MENU":GOTO935
965 PRINTDN$DN$
970 INPUT"ENTER GAIN OF XMIT ANTENNA (IN DB)";GX
975 INPUT"ENTER GAIN OF RCV ANTENNA (IN DB)";GR
980 PRINTDN$DN$"ENTER TOTAL LOSSES AT TRANSMITTING AND
985 PRINT"RECEIVING ENDS OF THE LINK.  INCLUDE
990 INPUT"COAX CABLE LOSSES, ETC. (IN DB)";LL
995 PRINTCL$:RETURN

9000 RD=7:PRINTDN$DN$DN$DN$,CHR$(1B)"READING DATA ";RD;
9005 DIM H$(2,3,5,9):DIMH$(2,3,5)
9010 FORPO=1TO2:REM 1=VERT 2=HORIZ
9015 FORPP=1TO3:REM 1=POOR SOIL 2=GOOD SOIL 3=SEA WATER
9020 FORFB=1TO5:REM 1=3.5 2=7 3=14 4=28 5=50MHZ
9025 FORAH=1TO9:REM 1=10' 2=20 3=30 ... 6=60 7=80 8=100 9=150'
9030 READ H$(PO,PP,FB,AH):REM HEIGHT GAIN IN DB
9035 NEXT:NEXT:RD=RD-1:PRINTLT$LT$LT$RD;:NEXT:NEXT
9040 DATA 0,0,0,0,0,0,0,0,0
9042 DATA -1,-1,0,.5,1,2,3,4,7
9044 DATA 0,1.5,3,5,7,8,10,12,15
9046 DATA 2,6,8,11,12.5,14,16.5,18.5,22
9048 DATA 6,11,14,17,18.5,20,22.5,24,28
9050 DATA 0,0,0,0,0,0,0,0,0
9052 DATA 0,0,0,0,0,0,0,0,0
9054 DATA -.5,-1,-1,-.5,0,1,2,4,7
9056 DATA 0,1,2,3.5,5,6,8,10,13
9058 DATA 1,4,7,9,11,13,15,17,20
9060 DATA 0,0,0,0,0,0,0,0,0
9062 DATA 0,0,0,0,0,0,0,0,0
9064 DATA 0,0,0,0,0,0,0,0,0
9066 DATA 0,0,0,0,0,0,0,0,0
9068 DATA -1,-1.5,-2,-2.5,-3,-3,-2,-1.5,1
9070 DATA 0,2,4,6,7,8,10.5,12,15.5
9072 DATA 0,3,6,7.5,9,11,13,15,18.5
9074 DATA 0,4,8,11,13,14,16,18,22
9076 DATA 0,5.5,9,12,14,15,17,19.5,23
```

```
9078 DATA 0,6,9.5,12.5,14.5,15.5,17.5,20,23.5
9080 DATA 0,4,7.5,10,11.5,13,16,18,21
9082 DATA 0,5,8,11,13,15,17,19,22
9084 DATA 0,6,10,12,14,16,18,20,23
9086 DATA 0,6,10,12,14,16,18,20,23
9088 DATA 0,6,10,12,14,16,18,20,23
9090 DATA 0,6,9.5,12,14,15.5,18,20,23
9092 DATA 0,6,9.5,12,14,15.5,18,20,23
9094 DATA 0,6,9.5,12,14,15.5,18,20,23
9096 DATA 0,6,9.5,12,14,15.5,18,20,23
9098 DATA 0,6,9.5,12,14,15.5,18,20,23
9100 H$(1,1,1)="09900710901211902012903513905414908815912 5
9102 H$(1,1,1)=H$(1,1,1)+"16917017921018925019929020932 0
9110 H$(1,1,2)="1054.51157.61250131350231450371550551650 88
9112 H$(1,1,2)=H$(1,1,2)+"17512018516019519020523021527 0
9120 H$(1,1,3)="1112.51214.51310081410131510231610371710 53
9122 H$(1,1,3)=H$(1,1,3)+"18108019110520113021150221180
9130 H$(1,1,4)="1181.81283.11385.51480101580161680251780 40
9132 H$(1,1,4)=H$(1,1,4)+"18805819807820809621812022813 0
9140 H$(1,1,5)="1231.31332.21430041530071630121730181830 30
9142 H$(1,1,5)=H$(1,1,5)+"19304320305821307522309023311 0
9150 H$(1,2,1)="09902710904511906012911013915014919015923 0
9152 H$(1,2,1)=H$(1,2,1)+"16927017931018934019938020942 0
9160 H$(1,2,2)="10501411502312503213506014509015512516516 0
9162 H$(1,2,2)=H$(1,2,2)+"17520018522519524020527021530 0

9170 H$(1,2,3)="1115.71210111310211410321510501610701710 94
9172 H$(1,2,3)=H$(1,2,3)+"18112019114020116021180221200
9180 H$(1,2,4)="1184.21287.51380131480221580351680501780 69
9182 H$(1,2,4)=H$(1,2,4)+"18808919811020813021815022817 0
9190 H$(1,2,5)="12300313300514301015301516302417303718305 0
9192 H$(1,2,5)=H$(1,2,5)+"19306720308321310022312023314 0
9200 H$(1,3,1)="09919010929011938012947013957014965015974 0
9202 H$(1,3,1)=H$(1,3,1)+"16982017992018999918999918999 9
9210 H$(1,3,2)="10515011521512528013535014542515550016558 0
9212 H$(1,3,2)=H$(1,3,2)+"17564018570019578020584021590 0
9220 H$(1,3,3)="11108312111513115014118515122016126017130 0
9222 H$(1,3,3)=H$(1,3,3)+"18134019137020141021145022148 0
9230 H$(1,3,4)="11803512805013807014809015811016813017815 0
9232 H$(1,3,4)=H$(1,3,4)+"18817019819020821021823022825 0
9240 H$(1,3,5)="12301713302714304015305416307017308518310 0
9242 H$(1,3,5)=H$(1,3,5)+"19311720313021315022317023318 5
9250 H$(2,1,1)="0991.51092.61194.81298.11390151490251590 42
9252 H$(2,1,1)=H$(2,1,1)+"16906017909018912519916220920 0
9260 H$(2,1,2)="1051.51152.61254.81358.11450151550251650 42
9262 H$(2,1,2)=H$(2,1,2)+"17506018508019511520514021517 0
9270 H$(2,1,3)="1111.51212.61314.81418.11510151610251710 42
9272 H$(2,1,3)=H$(2,1,3)+"18105619107820110021122221145
9280 H$(2,1,4)="1181.81283.21385.91480101580161680301780 40
9282 H$(2,1,4)=H$(2,1,4)+"18806019808020809621811622813 6
9290 H$(2,1,5)="1232.41334.11437.21530131630221730321830 45
9292 H$(2,1,5)=H$(2,1,5)+"19306020307121308223100233120
9300 H$(2,2,1)="0990.71091.31192.31290413900714901315902 2
9302 H$(2,2,1)=H$(2,2,1)+"16903617905518908019911020915 0
9310 H$(2,2,2)="1050.91151.61252.91350051458.81550161650 26
9312 H$(2,2,2)=H$(2,2,2)+"17504218506619509020512021515 0
9320 H$(2,2,3)="1111.31212.11313041410071510131610211710 35
9322 H$(2,2,3)=H$(2,2,3)+"18105019107520109521120221142
9330 H$(2,2,4)="1181.81283.21385.21489.51580161680261780 40
9332 H$(2,2,4)=H$(2,2,4)+"18805819807620809421811622813 5
9340 H$(2,2,5)="1232.41334.21437.41530121630201730311830 45
9342 H$(2,2,5)=H$(2,2,5)+"19305820307421309022311023312 0
9350 H$(2,3,1)="0990.61091.01191.71290031395.51499.5159017
9352 H$(2,3,1)=H$(2,3,1)+"16902817904618907219910820914 0
9360 H$(2,3,2)="1050.81151.41252.41354.31457.71550131650 23
9362 H$(2,3,2)=H$(2,3,2)+"17503818505919508420511521515 0
9370 H$(2,3,3)="1111.21212.11313.81416.81510121610201710 33
9372 H$(2,3,3)=H$(2,3,3)+"18105019107020109521120221140
9380 H$(2,3,4)="1181.71283.01385.41489.41580161680261780 40
9382 H$(2,3,4)=H$(2,3,4)+"18805619807620809821811622813 0
9390 H$(2,3,5)="1232.41334.11437.21530121630201730213045
9392 H$(2,3,5)=H$(2,3,5)+"19306020307621309022311023312 5
9400 F(1)=3.5:F(2)=7:F(3)=14:F(4)=28:F(5)=50
9405 RD=RD-1:PRINTLT$LT$LT$RD
9410 P$$(1)="VERTICAL":P$$(2)="HORIZONTAL"
9420 PP$(1)="POOR SOIL":PP$(2)="GOOD SOIL":PP$(3)="SEA WATER"
9430 LL$="----------------------------------------"
9440 DIMP(12):DIMD(12)
9450 H(0)=0:H(1)=10:H(2)=20:H(3)=30:H(4)=40:H(5)=50
9460 H(6)=60:H(7)=80:H(8)=100:H(9)=150
9999 RETURN
```


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gram assume vertical antennas are at ground level and horizontal antennas are at 10 feet (3.04 meters). In **lines 245-295** all your input parameters are displayed on the screen (see **fig. 3**). In **lines 300-335** a check is made to ensure that the results of your parameters fall within the dynamic range of the program, the expected communications range is calculated based upon the nearest two data points stored, and your expected communications range is displayed on the screen.

One of the features of this program is that you can experiment with changes in a single parameter without having to re-enter all the inputs. **Lines 400-426** print a menu at the bottom of the screen (see **fig. 3**). Would you like to see how much further you could communicate if you increased your transmitter power from 100 to 500 watts? Press "x" on the keyboard, enter new receiver sensitivity and transmitter power, and your new range will instantly be displayed. What about raising that antenna from 30 to 60 feet (from 9.14 to 18.28 meters)? Press "A," answer the questions, and your answer immediately appears. If you have a printer connected, just press "P" and **lines 430-436** will give you a screen dump for a hard copy of all the information.

data format

The propagation curves from which the data in this program were taken assume that both receiving and transmitting antennas are dipoles at 10 feet (3.04 meters) for horizontal polarization and grounded antennas for vertical polarization. In many cases for vertical antennas and in all cases for horizontal antennas, as you raise the height of the antennas, you can communicate over larger distances.

Each antenna elevation has an associated "height gain." For example, for horizontal polarization over *good soil* at 14 MHz, raising the antenna from 10 feet to 30 feet (from 3.04 to 9.14 meters) will give you the same increase in range as increasing your power from 100 watts to 1 kW or as increasing your antenna gain by 10 dB. Hence, a 30 foot (9.14 meter) antenna under these circumstances is considered to have a relative "height gain" of 10 dB.

The data statements in **lines 9040-9098** contain height gains for the various combinations of frequency, polarization, and soil type in this program. The first five lines are for vertical polarization over poor soil, the next five are for vertical polarization over good soil, etc., and the last five are for horizontal polarization over sea water. Each line contains nine numbers which are the height gains in dB for each of the nine discrete heights in the menu. For example, **line 9084** contains data for horizontal polarization over good soil at 14 MHz. The height gain for 10 feet (3.04 meters) is 0 dB (reference height); for 20 feet (6.09 meters), it's 6 dB; for 30 feet (9.14 meters), it's 10 dB . . . and for 150 feet (45.72 meters), it's **23** dB. These data are read into the four-dimensional array HG in **lines 9005-9035** during program initialization, and the proper height gains for the antenna condition you choose in the program are selected in **lines 230** and **235**.

There are 30 pairs of lines from **9100-9392**, one for each combination of polarization (two kinds), soil type (three kinds) and frequency (five bands). **H$(1,1,1)** contains data for vertical polarization over poor soil at 3.5 MHz; **H$(1,1,2)** is for the same at 7 MHz . . . through **H$(2,3,5)**, which is for the second polarization type (horizontal), the third path type (sea water), and the fifth frequency band (50 MHz). Each nonlinear propagation curve of path in dB vs distance in miles is broken into 12 data points, each consisting of three digits for path in dB followed by three digits for distance in miles for that point, and is reconstructed as an 11-segment "piece-wise" linear equation by the program. For example, the first six digits of **H$(1,1,1)** indicate that for 3.5 MHz vertical polarization over poor soil, a path of 99 dB will yield a communications range of 7 miles (11 km). Similarly, the last six characters of **H$(2,3,5)** predict a range of 125 miles (201 km) for a 233-dB path for horizontal polarization over sea water at 50 MHz.

The proper data set for the combination of frequency, path, and polarization chosen is selected in **line 210** of the main program. This is then broken into 12 path and 12 distance values by the character string manipulation in **lines 215** and **220**. In **lines 300-305**, your system path is checked to be sure it's within the program range: if not, a "range not in program" message is printed. In **lines 320-330** your path is compared with each of the 12 path points stored. When it's found to be between the two closest path points, the variable DI is calculated to be the same proportional distance between the associated distance data, and that distance is printed to the screen in **line 335**.

entering the program

Enter the program as listed, taking the normal precautions to SAVE it before you RUN it so that if you make a typing error that could cause a computer lock-up, you'll be able to go back to edit the saved version without having to retype the entire program. The remarks (REM...) are to make the program easy to follow and change. They don't need to be entered.

If you have a Commodore 64 or 128PC, the program will run as LISTed. If you have another brand of computer, you'll want to delete **lines 22-24** and **32-36**, which are machine-specific (or add your own color commands), and change the "SYS65126" in **line 416** to "STOP" or "END." These are the only commands in the program that are machine specific, so the program should be easily converted to run on most com-

puters using BASIC. Some changes might be necessary; for example, some computers won't permit multiple statements on a single line, so new line numbers will have to be added. The program as written requires just under 16K of RAM, but it could be "crunched" by deleting the title screen and REMarks statements.

If you don't want to keystroke the Commodore 64 program yourself, send a check or money order for $8.00 to me at the address shown at the beginning of the article. I'll send you a verified disk containing two copies of the program (1541 format). Add $2.00 for tape or if you live outside the United States. I also have an IBM-PC version available on disc for the same price.)

expected results

In free space, path attenuation changes at a 20 log(F) and a 20 log(D) rate where F is frequency in MHz and D is distance in miles. This means you need four times your present power (or a corresponding 6 dB increase in antenna gain) to double your distance because your radiated power is expanding over an increasing area which is proportional to distance squared. This also means that when your frequency increases by a factor of 10, the signal decreases 20 dB. The atmosphere doesn't really absorb the higher frequencies. A given radiated power will provide the same free-space field intensity (in microvolts per meter or watts per meter squared) at 30 MHz as it does at 3 MHz, for example, but a half-wavelength dipole at ten times the frequency is physically only one-tenth as long and therefore will "capture" only one-tenth as many microvolts (-20dB).

Although free-space attenuation is very predictable [loss in dB = 37 + 20 log(F) + 20 log(D)], actual attenuation is much greater due to the earth's curvature. In addition, some drastic and sometimes unexpected results occur when your antenna is within a few wavelengths of ground — which is usually the case below 30 MHz.

If you're using a vertical antenna at ground level, you can think of your system as if a return current were flowing through the ground, much as return currents flow through the shield of a coaxial cable. If the soil has low conductivity, the signal will be attenuated. For example, the same system that gives you a range of 125 miles (201 km) using a grounded vertical antenna at 3.5 MHz over poor soil will give you a range of 230 miles (370 km) over good soil and 740 miles (1190 km) over sea water. In general, your communications range with vertical polarization will be poor over poor soil and excellent over sea water.

If you place a horizontal dipole close to the ground (in terms of wavelengths), the earth will tend to act as a reflector. Because much of your signal will go "straight up," little will propagate along the horizon; the higher the conductivity of the earth, the more drastic the effect. For the same system parameters at 3.5 MHz referenced in the previous paragraph, if you have a horizontal dipole mounted at only 10 feet (3.04 meters) above ground level, your range will be about 42 miles (67 km) over poor soil, 22 miles (35 km) over good soil, and only 17 miles (27 km) over sea water.

Although the previous two paragraphs might lead you to beleive that vertical polarization is preferable to horizontal polarization for ground wave conditions (it sometimes is), other items should be considered. For example, a horizontal dipole at 3.5 MHz will be very efficient, but a vertical over a poor ground radial system may be less than 10 percent efficient. In addition, as a horizontal antenna is elevated to greater heights, the effects of ground described above will greatly decrease. Raising your 3.5 MHz vertical antenna from ground level to 100 feet (30.48 meters), for example, will have a negligible effect on your expected range. However, moving your horizontal dipole from 10 feet to 100 feet (from 3.04 to 30.48 meters) at this frequency will give "height gains" of 12, 18, and 20 dB over poor soil, good soil, and sea water, respectively.

At the higher frequencies, where antennas are generally mounted several wavelengths above ground, the propagation path is nearly independent of polarization. For example, at 50 MHz, using radio parameters of 100 watts out and 1 microvolt sensitivity, and assuming dipole antennas at 70 feet (21.33 meters) — with no losses for this "ideal" example — the predicted communications ranges fall between 54 and 60 miles (87 to 96 km), no matter what combinations of polarization and soil type you choose.

using the program

No more needs to be said about the mechanics of using the program. Some explanations and precautions, however, are in order.

1. Selection of frequency band, polarization, path type, and antenna height should be straightforward.
2. For transmitter power, use actual output power, not rated input power.
3. For receiver sensitivity, the actual receiver sensitivity can generally be used at the higher frequencies if you live in a "quiet" (QRN-wise) location. However, if you're operating 40 meters, where the background QRM and QRN levels are S6 or S9 or worse, you're kidding yourself if you use 0.2 microvolts, because you might not hear anything below 10 microvolts or so. Disconnect your antenna and listen to the noise level, then connect the antenna. If the background noise doesn't rise, use your receiver sensitivity. If the noise rises, use the background noise level (you can calibrate your S-meter with a signal generator).

4. If you'll be communicating over earth with low conductivity and low dielectric constant (such as dry sand, gravel, or rock), select "poor soil" from the propagation path menu. If you'll be communicating over soil with relatively high conductivity and high dielectric constant (such as cultivated farmland), select "good soil." Maritime mobile operators will obviously select sea water.

5. Antenna gain for horizontal polarization will generally be 0 dB for a dipole or inverted "V." If you're using a beam or directional array, enter the gain (in dB) relative to a dipole.

6. The curves for vertical polarization at ground level assume a grounded whip over a perfect ground radial system. If you have a phased array, enter the gain. If you have a moderate or poor ground radial system (most hams do), performance will be poor. Even with 15 ground radials, efficiencies as low as 50 percent can be expected.[4] For this case, your gain would either be −3 dB, or 0 dB with 3 dB added to other system losses. If you have only a ground rod or a few short radials, your efficiency may be less than 10 percent (−10 dB).

7. The height gains for elevated vertical antennas assume a whip over a good counterpoise (such as a ground plane antenna). Although the theoretical gain of a ground plane antenna is 3 dB above a dipole, the pattern maximum is normally about 30 degrees above the horizon, and the signal is usually down about 6 dB at the horizon (or 3 dB below a dipole level). If you're using a vertical dipole or beam antenna elevated more than a wavelength above ground (such as a 6-meter beam above 20 feet, or 6.09 meters), add 3 dB to the antenna gain. However, the presence of conducting tower and coaxial cables parallel to the driven element usually distort the radiation pattern, so expect actual performance for vertical antennas to be less than predicted.

8. Remember to include losses at both the receiving and transmitting ends of the circuit for total systems losses.

9. As with any "prediction" program, use the results with caution. The curves are based on propagation over average terrain under average conditons. Actual variations of 20 dB or more can be expected. Neither "poor" conditions nor "extended ground wave" conditions will surprise an experienced operator. Expect inferior performance if your antenna is not "in the clear" (i.e. in a jungle, inside an apartment building, etc.) or if your path involves mountainous terrain. Perhaps the most useful purpose for this program is to experiment with antenna height or gain and transmitter power, etc., "on paper" (more accurately, "in RAM") to determine relative system improvements without having to spend a lot of time and money constructing hardware and running comparative tests.

If you have comments or questions, feel free to write to me, but enclose an SASE if you wish a response.

references

1. National Defense Research Committee 15, *Propagation Curves, Report 966-6C*, Bell Telephone Laboratories, Inc., Issue 3, October, 1944 (declassified to "open" status March 8, 1946).

2. *ESSA Technical Report ERL 111-ITS 79: Transmission Loss Atlas for Select Service Bands from 0.125 to 1.5 GHz*, Institute for Telecommunication Sciences, Boulder, Colorado, May, 1969. (Available for $1.25 from Superintendent of Documents, United States Government Printing Office, Washington, D.C. 20402.

3. Lynn A. Gerig, WA9GFR, and Joseph R. Hennell, "Trade off Power for Antenna Gain at VHF?," *ham radio*, July, 1985.

4. *The ARRL Antenna Book*, American Radio Relay League, Inc., Newington, Connecticut, 1964.

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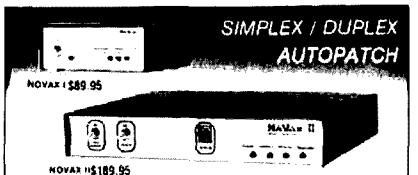
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how high
should your hf antenna be?

Knowing the desired range determines height

The question, "How high should my antenna be?" is often asked. But the only way to answer that question is by asking another: *how far do you want to communicate?* It's the answer to this question that allows the first to be answered.

Most hams build an antenna and then determine its capabilities and limitations by experiment. They never know whether poor communications should be blamed on the antenna or on propagation. A systems engineer, on the other hand, begins the process of designing a communications link by first determining the required distance for the link and then selecting the proper antenna, which is then built to the height that provides the optimum radiation pattern for the desired communications link. When link quality is poor, the designer can be certain that propagation conditions — not the antenna — are the limiting factor.

skip angle versus distance

The choice of antenna type and height can be readily determined from two simple graphs. First select the distance of the desired communications link. With that value in mind, refer to **fig. 1** to determine the optimum radiation angle or elevation take-off angle.

The graph shown in **fig. 1** is based on the average height of the ionosphere; although this distance varies, the typical Amateur antenna has a broad radiation pattern in the elevation plane, so high accuracy is not required, either for the height of the ionosphere or the height of the antenna.

Typically, we want to be able to select from a broad range of communications distances rather than be limited to only one location. So you'll need to determine the range of communications distances and the resulting range of optimum elevation radiation angles. If the range of radiation angles is too broad, more than one antenna may be required to optimize station performance, as will be shown later.

height versus skip angle

Figure 2 shows a computer-derived graph of antenna height versus skip angle. To allow the graph to be universal, antenna height is shown as a function of wavelength. Therefore, when you find the desired height, the physical height can be determined from the familiar equation of $\lambda = 984/f$, where λ is measured in feet and f is the operating frequency in MHz. For example, if the optimum height 0.75 λ, find λ for your operating frequency by multiplying it by 0.75. This allows the graph to be used at any frequency.

To explain the use of the graph, let us say we determined the distances we wanted to cover to be a range of 10 to 500 miles. From **fig. 1**, this translates to a radiation angle range of 87 to 37 degrees. From **fig. 2** we find the antenna could be anywhere betwen 0.1 and 0.4 λ for 87 degrees and between 0.2 and 0.6 λ for 37 degrees. To cover the desired range we need a horizontal antenna located between 0.2 and 0.4 λ. Since this is probably a good ragchew antenna for 75 meters, this translates to: $\lambda = 984/3.9 = 252$ feet (76.9 meters). Therefore, 0.2 λ = 50 feet (15.4 meters) and 0.4 λ = 101 feet (30.8 meters).

The black area of the curve covers the 3 dB beamwidth of the antenna. At 50 feet the antenna gain would be down 3 dB at a takeoff angle of 37 degrees, and down 3 dB at 87 degrees. The peak of the antenna pattern would be in the center of the black area of the curve.

Understanding the graph may be simplified by relating it to a "standard" antenna pattern. **Figure 3** shows the pattern of a dipole 1.25 λ above ground. If you were to draw a line on **fig. 2** at a height of 1.25 λ, you'd note that it would cross the first black curve at

By Ted Hart, W5QJR, P.O. Box 334, Melbourne, Florida 32902

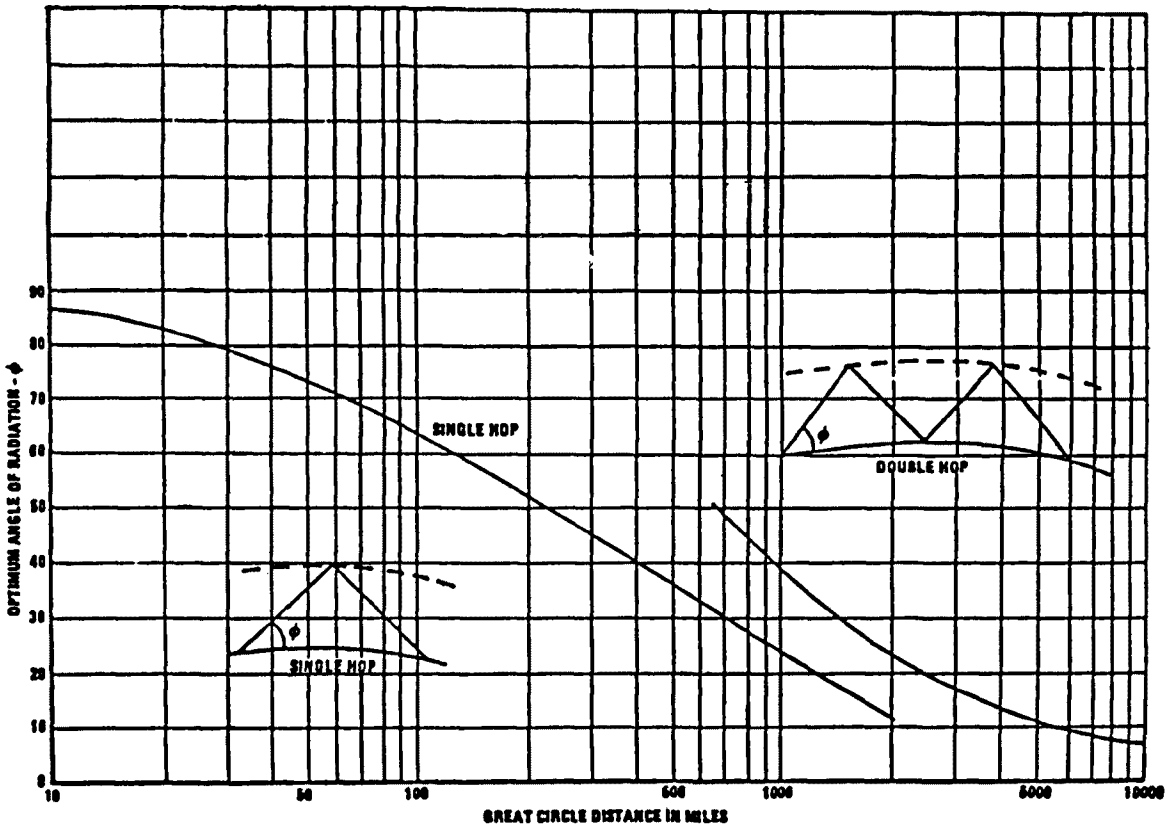


fig. 1. Optimum angle of radiation vs. distance for 1-hop and 2-hop communications based on average height of ionosphere.

5 degrees and 18 degrees (corresponding to the first lobe in the pattern), at 30 degrees and 45 degrees (corresponding to the second lobe), and at 62 degrees and continuing through 90 degrees on the third lobe. Hence, the multiple black areas correspond to the multiple lobes of the antenna pattern, when the antenna is high above ground. The areas between the black areas correspond to the nulls in the antenna pattern.

For a beam antenna, the same curves apply, except that the beam reduces the lobes on the back side because of the front-to-back ratio.

The graph presents only the elevation angles of the various lobes in the antenna, not the relative amplitudes. Typically, one lobe will be predominant in amplitude, with the other lobes at a reduced level. Since very few ham antennas are used at heights above one wavelength, this is a secondary consideration.

The graph also applies to vertical antennas. Normally, Amateurs use verticals only at ground level, but if the vertical were raised to a great height, multiple lobes wold appear in the radiation pattern. Since a vertical is a complement to a horizontal antenna, where one has a lobe the other would have a null in its pattern, asssuming both were at the same height. When using the graph for vertical antennas, simply use the areas that are not dark to derive the pattern.

what are your antenna's characteristics?

By working backwards, you can readily determine the range of communications for your existing antenna. Measure the height of the antenna and draw a corresponding line on **fig. 2.** Determine the elevation angles of the lobe(s), transfer that information to **fig. 1**, and read the corresponding range of communications distances. Again note in particular the nulls in the antenna pattern and the corresponding range. Now you know why you rarely talk to anyone at a distance that corresponds to a null in the antenna pattern. Also note that horizontal antennas close to the ground do not provide signals at low angles; consequently they're not useful for long range communications.* Conversely, vertical antennas don't radiate at high angles and are therefore not useful for short-

*Unless other modes of propagation exist — e.g. ducting, M, N, derived, etc. — Ed.

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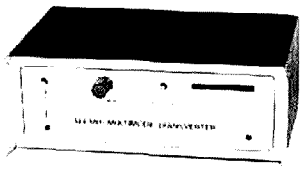
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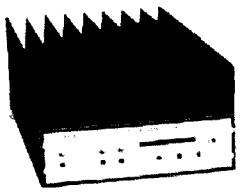
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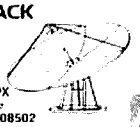
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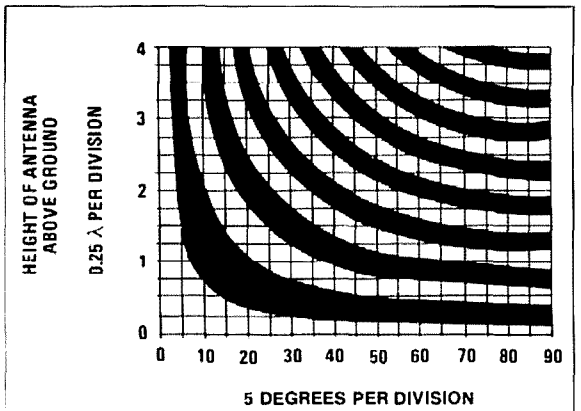
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fig. 2. Antenna height vs. skip angle.

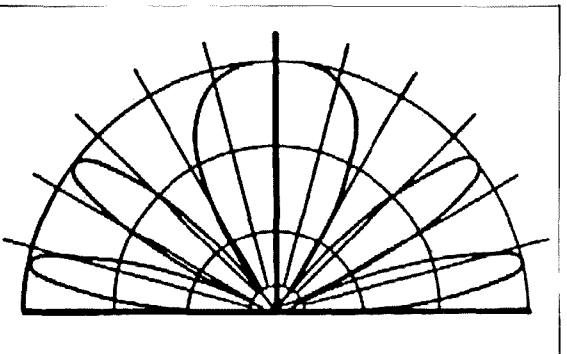


fig. 3. Dipole 1.25λ above ground.

range communications (less than 500 miles). Truly complete Amateur stations have both dipoles and verticals for the low bands.

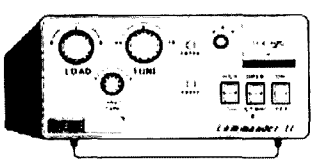
summary

Figures 1, 2, and **3** were reproduced from my book, *The Rules of The Antenna Game — Alias What Every Ham Must Know About Antennas*, available from the author for $5.95 + $1 for postage and handling. Please address inquiries to Ted Hart, W5QJR, W5QJR Antenna Products, P.O. Box 334, Melbourne, Florida 32902.

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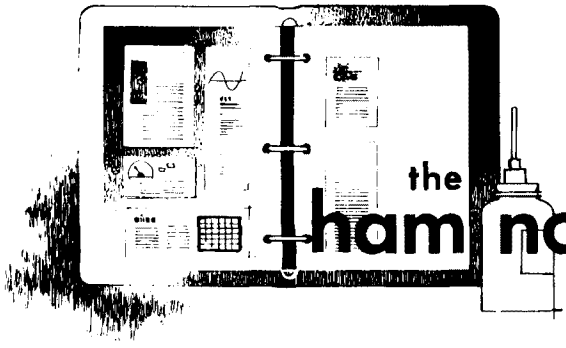
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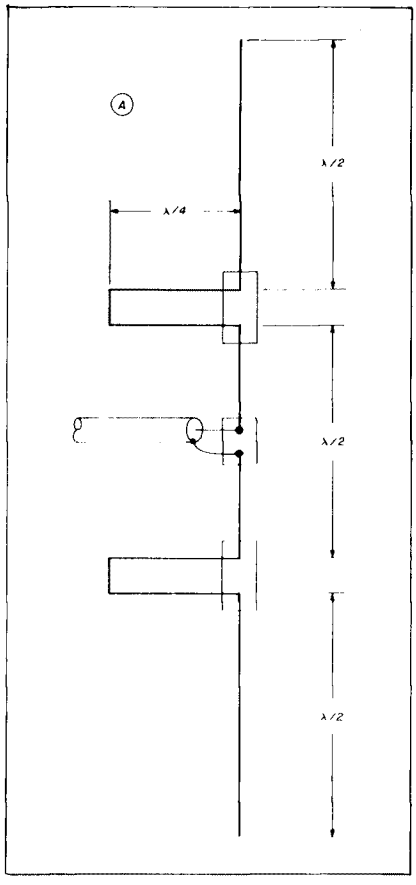
Depending on the size and construction techniques chosen, it's possible to build vertical or horizontal collinears in as little as one hour — or at most, one day. The cost is low, and gain can be increased by adding elements, one set at a time, to the array. Support can be a rope attached to a tree, a tower, or a building.

With only moderate variation in gain and SWR, a 2-meter collinear can be cut and matched for the middle of the band and it will work over the entire 4 MHz.

The antenna illustrated in **fig. 1** is a string of half-wave elements laid end-to-end and fed in phase by a series of half-wave delay lines (quarter wave shorted stubs). For use on 2 meters, the antenna is made from No. 12 solid copper wire. PVC pipe is used for insulators.

My antenna is fed with coax. The VSWR is less than 1.5:1 over the entire band at the transmitter. With a ten-element array, I've had contacts with mobile stations 50 miles (80 km) away and constant contacts through repeaters 80 to 100 miles (128 to 160 km) away.

If side-mounted on a tower or pole, three collinears can be phased so that two can be used simultaneously, with an additional 3 to 4 dB gain realized in any six directions.



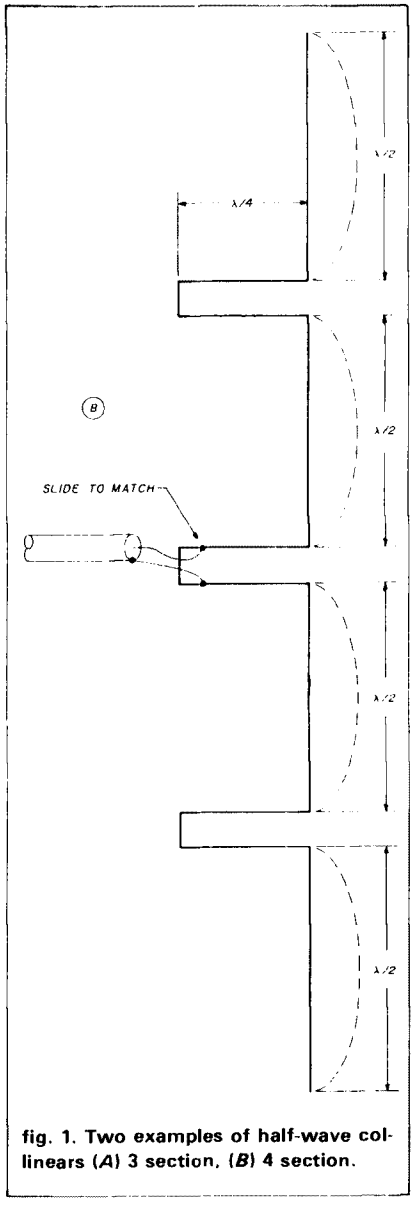


fig. 1. Two examples of half-wave collinears (*A*) 3 section, (*B*) 4 section.

gain over a dipole	
2-element	1.9 dB
3-element	3.2 dB
4-element	4.3 dB
10-element	10.0 dB
20-element	13.0 dB
40-element	16.0 dB

The phasing stub supports are PVC pipe measuring 2.5 inches (6.35 cm) long and drilled 0.5 inch (1.27 cm) from the ends. By changing the dimensions for other bands, this type of antenna can be used horizontally for gain on 432 MHz and 6, 10, and 15 meters, etc. On 6 meters it works especially well for auroral contacts when set up to radiate horizontally, north and south. It's better than a 5-element beam.

The feedpoint impedance is a function of the array.

Figure 2 provides the actual lengths used in a six-section collinear together with construction notes. Provide a loop at both ends for the support rope

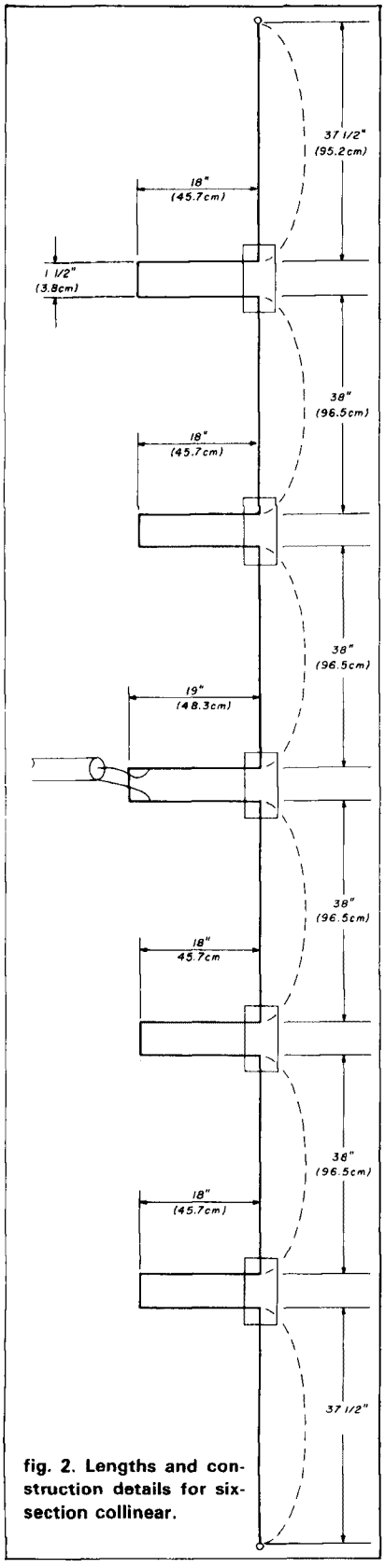


fig. 2. Lengths and construction details for six-section collinear.

or insulator and support wire. Assemble all driven elements, then add the phasing stubs, made from No. 12 or larger stiff copper wire, to be self-supporting for vertical mounting. Slide the coax feed along the driven stub in order to achieve the lowest VSWR. **Figure 3** shows one method of increasing the gain of the "Ringo Ranger" through the addition of a two-section collinear.

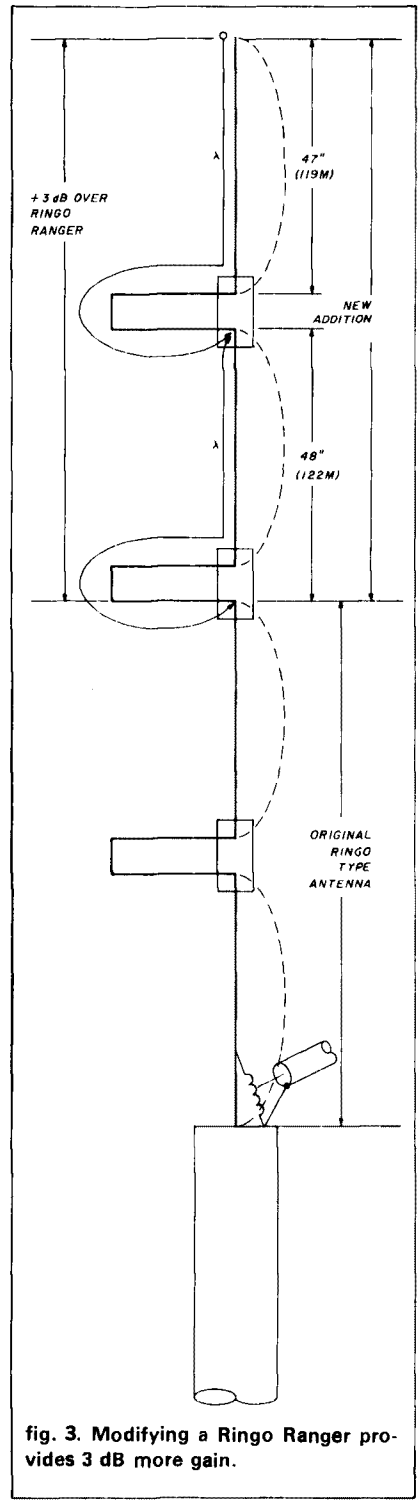


fig. 3. Modifying a Ringo Ranger provides 3 dB more gain.

larger design

I've built antennas using this design for 435 MHz, 439 MHz, and 2 meters for horizontal operation. My horizontal 2-meter antenna is a 40-element version in an inverted V form, made from No. 22 stranded teflon-coated wire.

acknowledgements

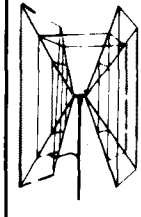
My thanks to WA1YJZ and WA3PGL for the tests and support while I experimented with these antennas over the years.

Please enclose an SASE with any inquiries, which should be addressed to me at R.D. 1, Union City, Pennsylvania 16438.

Leonard E. Shick, WB3AYW

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VHF/UHF WORLD

Joe Reisert W1JR

Yagi facts and fallacies

Surely the Yagi beam is the most commonly used Amateur antenna between 14 and 450 MHz — and rightfully so. It can yield higher gain for its physical size than any other antenna type. Its low profile makes it an excellent choice for those who are either stacking an array for very high gain or desire to operate several bands using the same antenna mast.

Over the past ten years there has been a definite improvement in the gain and especially the radiation pattern of long, high-gain Yagi beams. However, with these improvements there have also been seemingly unexplainable performance failures that have created confusion. Furthermore, pretentious gain claims have been made by some well-intentioned Amateurs.

In light of the above, I'm going to devote this month's column to the Yagi beam. Emphasis will be placed on attaining optimum performance. Certain problem areas will be discussed. I hope this material will be informative and put some of the myths to rest. In addition, it should provide guidance on how to select the proper parameters and obtain optimum performance when designing and building Yagis.

history

First, a few notes on the development of the Yagi-type antenna might be in order. The basic Yagi antenna structure as we know it today is classified as an "end fire" array. It usually consists of a single driven element with a reflector and one or more directors (**fig. 1**). The reflector is used mainly to decrease radiation off the rear of the antenna; the directors primarily determine the shape of the radiation pattern (or gain) in the forward direction.

The Yagi antenna was first developed in Japan during the late 1920s by Dr. H. Yagi and S. Uda.[1] Dr. Yagi published abroad in English, giving rise to the "singular" credit for this type of antenna.[2] His work with Uda on this antenna was published in book form in reference 3.

In the early days the Yagi antenna was typically used with only two to six elements. In the 1950s Kmosko and Johnson published one of the first really long Yagi antenna designs.[4] However, their heuristic ("cut and try") work had somewhat less gain than claimed and the radiation pattern had many sidelobes. These designs used tapered spacing and little or no director length tapering.

Greenblum published what was probably the first cookbook on Yagi designs.[5] However, how his graphs were to be used wasn't completely clear; some missed the point that the elements were through the boom, and the director lengths included the boom correction. The director spacings were not specific (a range of values was given) and some director lengths were also missing on the original charts. Regardlesss of the above, Tilton had great success using Greenblum's designs.[6,7,8]

Ehrenspeck and Poehler showed yet another design approach primarily aimed at decreasing the number of elements in the director structure; patterns were not optimized.[9]

Many years later, Dr. Ehrenspeck suggested that the best design for gain was probably the 0.3 wavelength spaced model.[10] He also said that if two additional directors were added to the front of the structure (as shown in **fig. 2**), that pattern could be improved and the gain increased by up to 1.0 dB.

Other individuals likewise published specific designs, but Amateurs really had to wait until Viezbickie and Reisert published the NBS Yagi designs.[11,12] Unfortunately, the NBS designs covered only six specific boom lengths, 0.4, 0.8, 1.2, 2.2, 3.2, and 4.2 wavelengths. Therefore, one couldn't easily design intermediate or longer boom designs (more on this later).

Hoch published design curves for Yagis of any length up to 47 directors.[13,14,15] His initial work was based on the designs of Greenblum.[5] He was able to improve the patterns and gain as well as account for the boom corrections. Hoch's designs work particularly well with booms 2 through 20 wavelengths long.

However, credit for the really serious work on tying down the optimum element lengths and spacings for Yagi antennas should probably go to Morris.[16] In his PhD thesis, he wrote computer programs to not only determine Yagi antenna patterns but also to optimize the lengths and spacing based on the desired pattern and gain.

Morris's work was followed by that of Chen and Cheng and others in the professional community who had access to large mainframe computers capable of handling the complex current matricies.[17,18] The late Dr. Jim

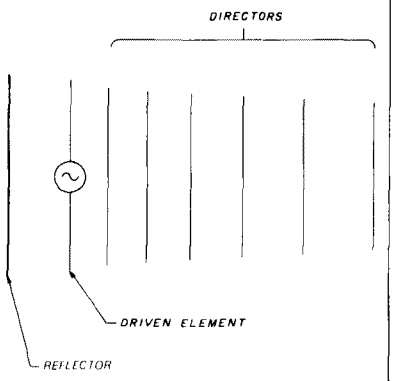


fig. 1. This is a typical Yagi antenna structure.

Lawson, W2PV, published similar work in the Amateur field, although his designs were limited to about 1.75 wavelengths.[19] Finally, Stan Jaffin, WB3BGU, published a program to do Yagi pattern measurement with a home computer,[20] giving the enterprising Amateur all the tools he or she needed to do serious Yagi design and optimization.

facts and fallacies

The above history is brief but necesary for understanding the development of the Yagi. Now let's proceed into the general area of facts and fallacies about it:

1. *The gain of a Yagi antenna is determined by the number of elements.* This is false. Gain is primarily a function of the length of the boom. The number of directors and their length and placement on the boom is a secondary factor that determines whether the gain for a particular boomlength is achieved, as well as the bandwidth, pattern shape and sidelobe levels.

Figure 3 is used to illustrate this point. This graph shows the highest possible gain that can be attained by a Yagi antenna on a specific boomlength (in terms of wavelengths). Note that the gain of a typical well-designed Yagi antenna increases in a logarithmic fashion at about 2.2 dB for each doubling of the boom length. Gain is shown both in dB over a dipole and dB over an isotropic.[23]

Therefore, using the graph in **fig. 3**, it can be shown that the highest possible gain for a 5-wavelength boom Yagi is approximately 15.3 dB over a dipole or 17.45 dB over an isotropic radiator. In reality, it will seldom be possible to attain this gain. If you can get within 0.5 dB of the value shown, consider yourself lucky!

2. *A greater number of directors for a specific boomlength can increase the bandwidth of a Yagi antenna and improve the radiation pattern.* This is true. There are a minimum number of elements needed for each boomlength. Evidence of this was described above in the discussion of the Ehrenspeck and Poehler Yagi designs.

However, if only this minimum number of elements is used, the pattern may have poor sidelobes, the front-to-back ratio may be low, and/or the frequency operating bandwidth narrow. You may ask why the latter item is important, since most weak signal operators operate only over a narrow bandwidth. The reason is that less pattern distortion will be prevalent with lesser element tolerances, weather changes, or structural changes — e.g., if an element loosens or breaks.

3. *Yagi antennas with high gain usually have good front-to-back ratios.* This isn't always true. The maximum gain and highest front-to-back ratio are not always coincidental for a specific boomlength. This means that for some boomlengths, either the gain or the front-to-back ratio may be optimum, but not necessarily coincident.

The NBS Yagis are an example of this phenomenon.[11,12] The 0.8 through 4.2 wavelength models were carefully designed to have both good gain and reasonable front-to-back ratio at the same time. Viezbickie noted that the gain increased similar to that shown in **fig. 3**, but small oscillations above and below the line were noted. He also stated in his report that certain boomlengths had slightly higher gain.

It has since been shown by computer analysis that certain boomlengths naturally exhibit high front-to-back ratio and optimum gain simultaneously.[19]

For the majority of moderate boomlength designs (less than 2 wavelengths), and especially the NBS designs, the optimum boomlength tends to be a multiple of odd quarter wavelengths long (for example, 0.75, 1.25, 1.75 wavelength, etc.). This probably explains why the particular NBS designs were chosen.

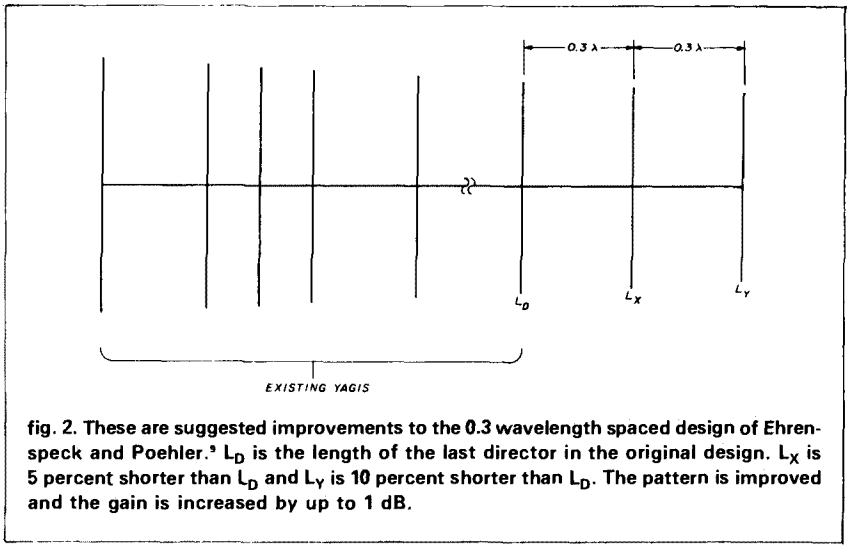


fig. 2. These are suggested improvements to the 0.3 wavelength spaced design of Ehrenspeck and Poehler.[5] L_D is the length of the last director in the original design. L_X is 5 percent shorter than L_D and L_Y is 10 percent shorter than L_D. The pattern is improved and the gain is increased by up to 1 dB.

At my request Stan Jaffin, WB3BGU, ran some examples of my DL6WU designs[15] on his mainframe computer program.[20] The results were quite interesting and I've plotted this data on the graph shown in **fig. 4**. Note that in the DL6WU designs there are also optimum boomlengths, but they seem to be spaced about 2 wavelengths apart. I would recommend using, if possible, the specific boomlengths where the front-to-back ratio is highest.

4. *Tubing and rods of the same diameter have the same electrical wavelength.* This is definitely not true, but this fact appears to have escaped most Yagi designers, especially at the lower frequencies, where tolerances are not as much a problem as they are in the UHF range. The NBS Yagi designs used rods for their elements.[11] Therefore, if you use an NBS design with tubing, the electrical length may not be the same as it would be with an equivalent rod. A UHF design may be slightly off frequency.

This phenomenon was apparently known by those who did the NBS Yagi designs. One document showed that if the end of the element was hemispherical instead of a flush cut, the frequency shifted upwards. The recommended change was to add an overall lengthening factor of 0.4 times the diameter to the element (0.2 times the element diameter to each end of the element).

Steve Powlishen, K1FO, has also noticed this phenomenon and has seen even a chamfer on the end of an element shift the frequency up perhaps 1 MHz or so at 432 MHz. A word to the wise: if you don't use solid rods and/or don't cut the end of the element off flush, check the resonance to see if the frequency shifted upwards!

5. *Scaling elements must be done using the published charts and graphs similar to those of Viezbickie, Reisert, and Hoch.*[11,12,14,15] This is not true. Scaling performed on these charts can be accurate if you don't make any mistakes in the process!

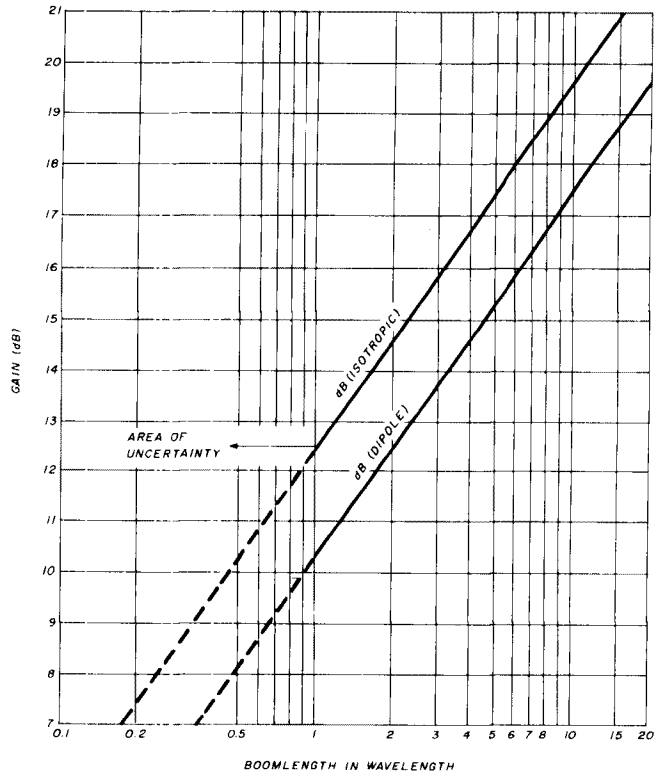


fig. 3. This graph represents the highest possible gain that can be achieved on a conventional (single plane) Yagi antenna. See text for further explanation.

However, scaling can also be done quickly — with probably greater accuracy — on a computer using the methods and equations 8 through 13 proposed by Lawson.[21] These equations can easily be programmed on a personal computer to yield rapid and accurate data without referring to graphs.[22]

6. *The NBS designs are the best Yagi designs available.* This is false. The NBS designs are good and reproduceable. However, they represent only six specific models, as discussed above.

Computer analysis has shown some discrepancies in these antenna designs.[19] Others have found that the patterns and gain don't match the data in the technical note at the design frequency on several of the models. Stan Jaffin, WB3BGU, has shown that if an extra director is placed approximately 0.15 wavelength ahead of the driven element on the 4.2 wavelength design, the gain can be increased by almost 0.5 dB.[20] This has been verified by at least one antenna manufacturer.

The NBS designs are good and reproduceable. If you feel comfortable with them, and if the boomlengths presented fit your needs, by all means use them. Perhaps you should also try the extra director. However, if you add this director to an existing antenna, it will change the impedance match. If the boomlengths of the NBS Yagis are too short, try one of the DL6WU designs.[15]

7. *The best reflector system is the trigonal method proposed by NBS.* This is definitely *not* true. When I first tried this reflector system on a 3.2 wavelength NBS Yagi, the gain dropped by almost 1.5 dB below the same antenna with a standard reflector. Repeated tests showed that the reflector lengths suggested by NBS were definitely too short.

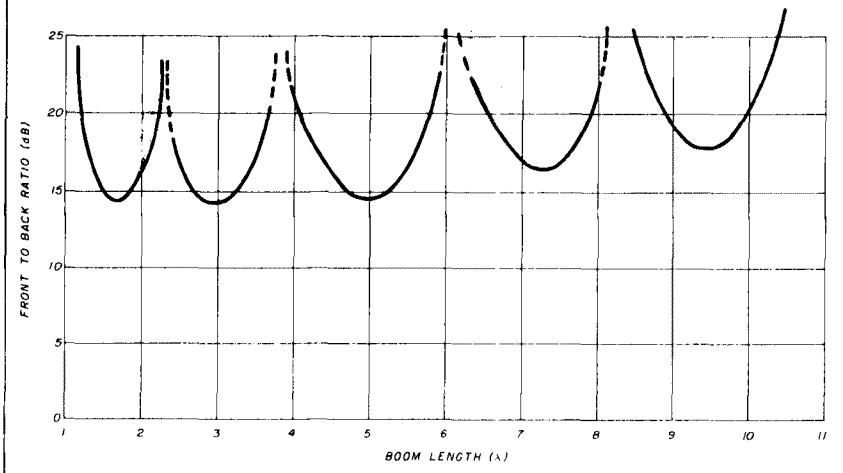


fig. 4. This is tabulated data from computer analyzed front-to-back ratio versus boom-length for the DL6WU Yagi designs.[15] Note the boom lengths where the front-to-back ratio is naturally highest.

I lengthened all three reflectors by 0.007 wavelength, the length needed to make them similar to the existing single reflector. Voila! The gain came back to normal. Repeated measurements showed, however, that the gain of a trigonal reflector system over a single reflector as suggested by NBS was only about 0.1-0.2 dB with perhaps 6 dB better front-to-back ratio. In hardware alone, this represents quite an increase in mechanics and wind load!

Several years later I mentioned this to Stan Jaffin, WB3BGU, and suggested that he could somehow test my theory on his computer program.[20] His tests confirmed my results and showed the same gain changes with the optimum lengthening factor to be 0.009 wavelengths.[24]

8. *The VSWR of an a Yagi is not important.* This is false. Although a moderate (2:1) VSWR would not seem to be important, it can be detrimental at VHF and especially UHF. The reason is that the feedline losses increase, especially if the nominal insertion loss is high (>1dB).

High VSWR can also be a sign that there is something wrong with the design. I once noticed high VSWR on a commercial antenna, only to find that the position of the hole for one of the directors was misdrilled. In addition, if the VSWR on an antenna is low and it changes, this can be an indication of trouble.

Then there's stacking, a typical way to increase gain on VHF and above. A high VSWR may have a very adverse effect when two or more antennas are summed together. Poor matching could divert more power to one of the antennas which would "hog" the power and thus decrease the anticipated gain increase.[25]

9. *Stacking antennas is a good way to increase gain.* This is true. However, the antennas must be properly designed and stacked a certain distance apart to obtain the increased gain. If the spacing distance is too close, the gain increase will be low. Stacking too far apart will increase sidelobes and noise pickup. The basics of stacking are thoroughly discussed in references 25 and 26.

10. *Elements that are insulated from the boom of a Yagi work better than those that are in ohmic contact with the boom.* This is definitely false. There are advantages and disadvantages to either type of mounting.

Insulated elements are less likely to induce boom resonances on other bands, a common problem on HF. While it can be argued that the insulators don't corrode, it can also be proved that the dielectric material in the insulator can get contaminated or deteriorate with age and exposure to the sun and weather.

Mounting elements in ohmic contact with a boom is a technique that has been around for a long time. This method is less likely to produce problems with static buildup and stray HF pickup from the feedline. Usually this technique is easier to use, and if the elements are properly installed without dissimilar materials, the corrosion problem is minimal.

Finally, some of the myths about element mounting have been perpetrated by those who say that antennas with insulated elements are detuned during wet weather. Tests have shown that by pouring water on an existing antenna with and without insulated elements that the detuning effects are about the same for either method *using a similar Yagi design.*

The primary reason for detuning in a Yagi antenna during wet weather is the sensitivity to element diameter. When ice or water is present on an element, its electrical length is changed. The higher the gain and the closer the antenna is operated to its cutoff frequency, the more the detuning effect will be noticed.

11. *Boom corrections are not important since the boom does not detune the element.* This is a serious misconception. The easiest way to envision the electrical characteristics of a boom is that it shorts out part of the elements. Therefore, any elements passing near or through a boom must be *lengthened* to reestablish the intended electrical length.

Fortunately, if too little correction is applied, the frequency of a Yagi is increased. It is well known that a Yagi antenna has a very rapid cutoff above resonance and a slow cutoff below resonance. Hence, if the correction factor is too small, only a slight degradation in performance will be noticed.

Boom corrections seem to be a big

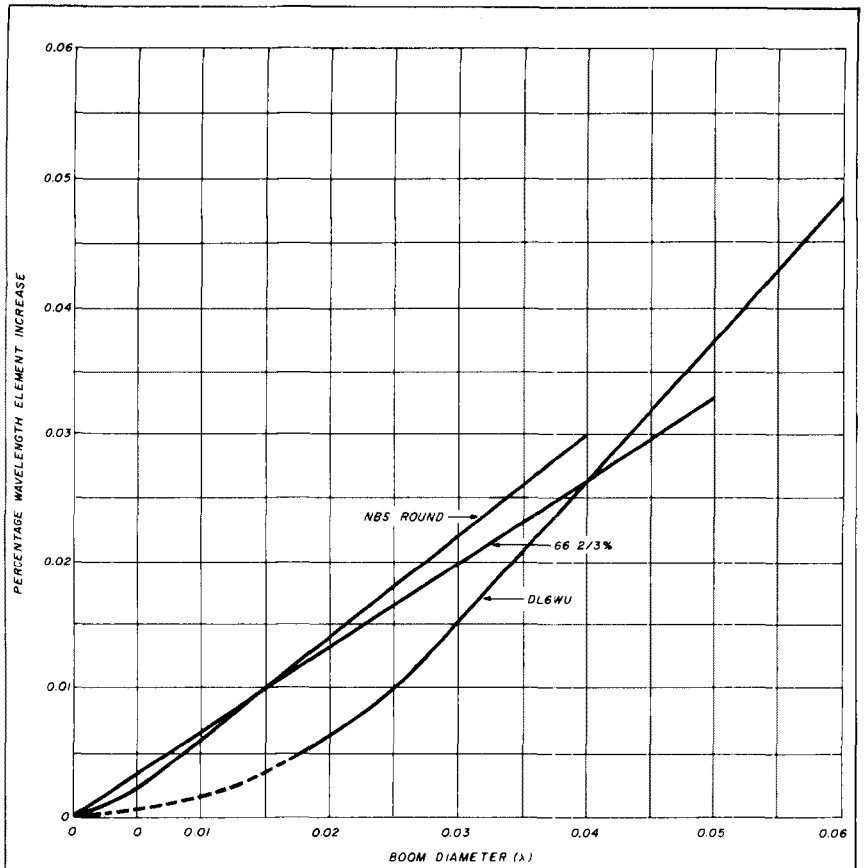


fig. 5. This graph shows three of the most popular boom correction factors. See text for further explanation.

mystery despite the fact that they've been mentioned in the literature for many years. A correction must be made whenever an element passes through or within one radius of a boom with a diameter exceeding 0.0025 wavelengths with respect to the operating frequency (0.2 inches or 5 mm at 144 MHz).

Several different correction factors have been considered. Many years ago a 66 percent correction was recommended for through the boom elements.[5] Later we had the NBS corrections.[11,12] Then DL6WU proposed a correction factor.[13] Unfortunately they don't all agree, but they are close! I have shown these corrections in **fig. 5.**

What if you don't go through the center of the boom or use insulated elements? I have measured some effects on my "backyard" antenna range. For mounting elements on top of the boom a la Cushcraft, the correction seems to be about 0.32 inches (8 mm). Using through the boom insulated elements such as the K2RIW 19-element 432-MHz Yagi seems to be about 40 to 50 percent of the NBS corrections.

Finally, when mounting above the boom, I constructed a sort of pyramid correction factor shown in **fig. 6**. It can be used to estimate the percentage of change based on where the element is mounted with reference to the boom center. For instance, if the element is mounted at least one boom radius above the boom there is no correction, but mounting right on top of the boom would require about a 50 percent correction as opposed to mounting directly through the middle of the boom.

12. *The material used for Yagi elements is important.* This is true. Aluminum is preferred since it's easy to work with, light in weight, and very efficient. Furthermore, aluminum-to-aluminum contacts are recommended. Brass is also usable, but it gets quite brittle after exposure to the weather. Copper is usually too expensive, heavy, and soft. Stainless steel elements, however, are not recommended since the skin effect will definitely lower antenna efficiency, especially above 1000 MHz.

13. *There is nothing wrong with changing an existing design.* This is definitely false unless you really know what you're doing. For example, some Amateurs have tried to second-guess the NBS designs and have invented new variations. A 4.2 wavelength NBS Yagi model at 432 MHz is less than 10 feet (3 meters) long, while standard tubing comes in 12-foot (3.5 meter) lengths. Why throw away the extra 2 feet (61 cm) of tubing, they ask? Just add on a few more elements to the extra tubing.

This approach has always resulted in disaster. The element lengths and spacings for each specific design are carefully chosen to yield a certain phase velocity. Changing a design calls for a new phase velocity. Failing to obtain the correct parameters, or changing elements arbitrarily, will usually decrease gain and distort the antenna radiation pattern.

What this boils down to is the following: if you must redesign an existing design, you can do so only if you have the proper tools at your disposal. These include, but aren't limited to, lots of know-how, an acceptable antenna range, perhaps a computer program, and lots of time and patience. Better yet, start out from scratch, using one of the NBS or DL6WU designs.

14. *Impedance matching is easy.* This is true, but you must know what you're doing. For instance, the length of the driven element in a Yagi isn't critical, but the matching method may be.

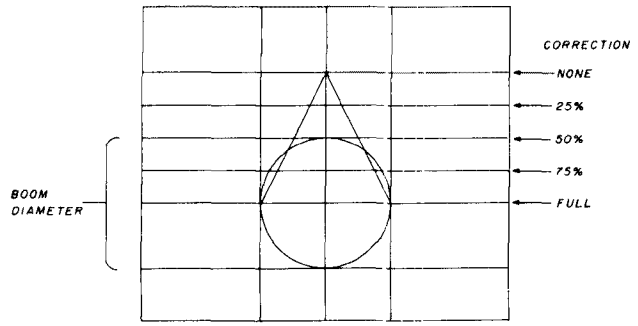


fig. 6. This figure will allow you to approximate boom corrections if an element is not placed directly through the center of a boom. (See text.)

It's been pointed out many times in this column that the Gamma match is especially poor above 150 MHz. Furthermore, Gamma matches often introduce unbalance into the antenna which may cause radiation on the feedline. A folded dipole with a 4:1 half-wave balun is an acceptable feed, but it can be difficult to match if the feed point isn't the proper impedance. I personally prefer the "T" match (without series capacitors) with a built in half-wave balun. It's inherently balanced, easy to tune, and efficient; it also suppresses the possibility of feedline radiation.

15. *Computer-designed Yagi antennas are coming.* This is true. First you'll have to have a computer program. Next, you'll need some antenna savvy on what to do to make a particular Yagi design work.

The NBS and DL6WU designs can be used for starters. Just pencil up a Yagi design. Then tweak the element lengths and spacing on the computer until you get the desired pattern.

I predict that before long, computer designs will be the most promising thing to happen in Yagi design in a long time. You'll be hearing more about this subject in the future. If you're so inclined, review references 19 and 20.

final evaluation

Now comes the fun. You've built that new super-high- gain Yagi and want to know if it plays. First the VSWR has to be matched; the lower the VSWR, the better — but 1.2:1 is more than sufficient.

If you have a radiation plot, your work is easy. Just measure the radiation pattern by the methods described in reference 23 and compare your results to the measured results. If the beamwidth is near the expected value and the sidelobes are down as many dBs as expected, you're probably in good shape.

If the beamwidth is too wide, the sidelobes are better than expected, and/or the first nulls are deeper than expected, the antenna may be tuned too high in frequency. If the beamwidth is too narrow, the sidelobes are worse than expected and/or the first nulls are shallow, the antenna is tuned too low in frequency. You're now on your way.

summary

This month's column was primarily aimed at taking the mystery out of Yagi antenna design and trying to dispel some myths. For those who are timid, the NBS[11,12] or the DL6WU[15] Yagi designs are recommended. If you have a personal computer and can obtain a Yagi program such as MININEC or those mentioned in references 19 and 20, you can "roll your own" without even cutting a piece of tubing!

acknowledgements

I'd particularly like to thank Günter Hoch, DL6WU, Stan Jaffin, WB3BGU, Steve Powlishen, K1FO, and Dr. Hermann Ehrenspeck for their discussions with me about some of the material presented in this month's column.

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25. Joe Reisert, W1JR, "VHF/UHF World: Stacking Antennas, Part 2," *ham radio*, May, 1985, page 95.
26. Joe Reisert, W1JR, "VHF/UHF World: Stacking Antennas, Part 1," *ham radio*, April, 1985, page 129.

ham radio

new records?

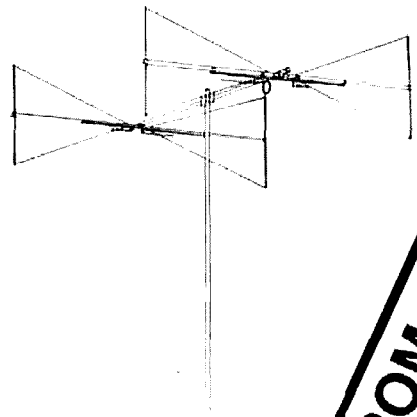
One of the greatest magnetic storms in recorded history occurred on February 8 and 9, 1986. While the HF bands went dead, numerous auroral contacts were made as high as 432 MHz over some incredible distances. If you had an aurora QSO over a longer distance than shown in **table** 4 of this column in the July, 1985 issue, I'd like to hear from you so we can see whether you've set a new aurora record. Forms for authentication are available for an SASE.

A new 9-cm (3456 MHz) world record was made by VK5QR and VK6WG across the great Australian Bight for a distance of approximately 736 miles (1185 km) on January 25, 1986. Soon afterwards, a new North American 13-cm (2304 MHz) overland tropo DX record was made between W4ODW and WB5LUA. Hearty congratulations to all! (Stay tuned for more details.)

important VHF/UHF events:

May 4:	*Predicted peak of the Eta Aquarids meteor shower at 1900 UTC*
May 8:	*ARRL 1296-MHz Sprint Contest*
May 10/11:	*Southern California 6-Meter Club QSO Party (contact N6FSL)*
May 16-18:	*12th Annual Eastern VHF/UHF Conference, Nashua, NH (contact W1EJ)*
May 17:	*ARRL 50-MHz Sprint Contest*
May 24:	*EME perigee*
June 6:	*Predicted peak of the daytime Arietids meteor shower at 0100 UTC*
June 7-8:	*ARRL VHF QSO party*
June 15:	*Predicted peak of the June Lyrids meteor shower at 2100 UTC.*
June 21:	*EME perigee*
June 21:	*Mean date* ±one month for peak of Sporadic-E propagation.

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A Compact Beam
for 20-15-12-10 Meters



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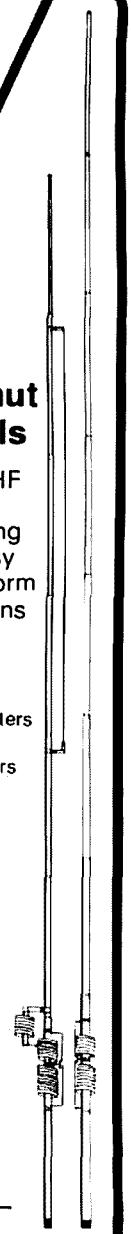
Model HF6V

- 80, 40, 30, 20 15 and 10 meters automatic bandswitching.
- Add-on kit for 17 and 12 meters available now.
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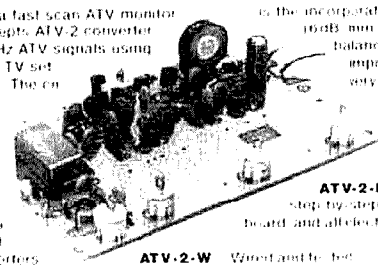
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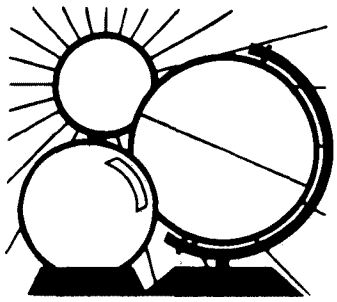
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DX FORECASTER

Garth Stonehocker, KØRYW

sporadic-E propagation

From May through September the overhead sun fills the lower ionosphere with ions that support short-skip propagation, even multiple short skips. The geomagnetic field clusters these ions into cloud-like patches known as *sporadic-E.* These patches form a thin layer of intense ionization about 60 miles (100 km) above the Earth.

A patch gives a strong, mirror-like signal reflection over skip distances of 600 to 1200 miles (1000 to 2000 km). Signals remain strong for about half an hour, up to a couple of hours after the onset of the first strong signal.

Station location determines how strongly the sunspot/geomagnetic disturbances affect sporadic-E propagation, with mid-latitudes the least affected and equatorial and polar paths the most affected. The best locations for these E_s openings are in the Northern Hemisphere from June through September and in the Southern Hemisphere during their summer, December through March. The best E_s is on either side of the geomagnetic equator; it's especially good where the geomagnetic equator is furthest from the geographic equator. These special areas are Southeast Asia in the Northern Hemisphere and South America in the Southern Hemisphere, with the former the better of the two.

The highest frequency propagated by E_s occurs at local noon, following the sun across the sky. However, the highest probability of occurrence is near sunrise and again around sunset. These two characteristics of E_s affect short-skip openings differently. Openings on the higher-frequency bands occur around local noontime; the lower bands tend to have openings near sunrise and sunset.

Most of us don't live in the special areas. The maximum E_s frequencies around the mid latitudes are 8 to 9 MHz. With the oblique factor of 5 for a 2000 km maximum-hop, the E_s-MUF becomes 40 to 45 MHz (almost 6 meters). So while the 10 and 6 meter bands have a good probability of opening up, 2 meter openings are rare indeed.

last minute forecast

The second and third weeks of May are expected to be the best for the higher frequency daytime bands, 10 to 30 meters. The solar flux 27-day variation should be maximum during those weeks. The lower frequency bands for daylight, short-skip, and nighttime DX are expected to be best the first and last weeks. Disturbed periods are possible around the 10th and 21st, with MUFs down 15 to 20 percent on the first day of the disturbance.

Of interest to moonbounce DXers, the lunar perigee occurs on the 24th. The full moon occurs on both the 1st and 30th of this month. The Aquarid meteor shower, of interest to meteor-scatter and meteor-burst DXers, peaks between May 4th and 6th, with rates of 10 to 25 per hour for the northern and southern hemispheres, respectively.

band-by-band summary

Six meters will provide occasional openings to South Africa and South America around local noontime by short-skip E_s.

Ten meters will be open to the southeast for a short period before local noon; to the south at noon and to the southwest after noon. Openings will last longer when the solar flux is at a maximum.

Fifteen and twenty meters, almost always open to some part of the world, will be the main daytime DX bands. Twenty meters should stay open on long southern paths into the night, though 15 will drop out in the late afternoon. Operate on 15 first, then move down to 20 meters later. DX is 5000 to 7000 miles (8000 to 11,300 km) on these bands. There may be some one-long-hop transequatorial propagation.

Thirty and forty meters are both daytime and nighttime bands. Intermediate distance operation, 1000 to 1500 miles (1600 to 2400 km), in any direction, is considered daytime DX. Nighttime DX on these two bands may be expected to occur over greater distances than on 80 meters and, like 80, will follow the darkness path across the sky. Signal strength and distances covered are lower on days of high solar flux values. In addition, no 30-meter openings will take place during the predawn hours on the morning after these high radio flux values.

Eighty and one-sixty meters will exhibit short skip conditions during daylight hours and lengthen for DX near dark when the QRN isn't bad. Eighty meters will open to the east just before your sunset, swing more to the south as midnight approaches, and end up in the Pacific areas during the hour or so before dawn. (One-sixty opens later and ends earlier.)

GMT	PDT	WESTERN USA N	NE	E	SE	S	SW	W	NW	MDT	MID USA N	NE	E	SE	S	SW	W	NW	CDT	EDT	EASTERN USA N	NE	E	SE	S	SW	W	NW
0000	5:00	20*	30	20	12	15	10	10	15	6:00	20	20	20	12	15	10	10	15	7:00	8:00	20	20	20	12	20	10	10	15
0100	6:00	15	*30*	*20*	12	15	10	10	15	7:00	20	*30*	*20*	12	15	10	10	15	8:00	9:00	20	20	*20*	12	20	10	10	15
0200	7:00	15	*30*	*20*	15	20	10	10	15	8:00	20	*30*	*20*	15	20	10	10	15	9:00	10:00	20	30	*20*	15	20	10	10	20
0300	8:00	15	*30*	*20*	15	20	10	10	15	9:00	15	*30*	*20*	15	20	10	10	20	10:00	11:00	20	*30*	*20*	20	30	10	10	20
0400	9:00	20*	30	*30*	20	20	10	10	15	10:00	20*	20	*30*	20	20	10	10	20	11:00	12:00	*20*	*30*	*30*	20	30	12	12	20
0500	10:00	20	20	*20*	20	30	10	10	15	11:00	20	20	*20*	20	30	12	10	20	12:00	1:00	*20*	*30*	*20*	20	30	12	12	20
0000	11:00	20	20	15	20	30	12	10	20	12:00	20	*30*	20	20	30	12	12	20	1:00	2:00	*20*	*30*	20	20	30	15	15	20
0700	12:00	20	*20*	15	20	30	12	12	20	1:00	*30*	*30*	20	*20*	30	15	15	20	2:00	3:00	*20*	*30*	20	*20*	30	20	20*	*20*
0000	1:00	20	*30*	20	*20*	30	15	20*	20	2:00	*20*	*30*	20	*20*	30	20	20	*30*	3:00	4:00	20	20	20	*20*	40	20	20	*20*
0000	2:00	*20*	*30*	20	*20*	30	20*	20	20	3:00	*20*	20	20	*30*	40	20	20	*30*	4:00	5:00	20	20	20	*30*	40	20	20	20
1000	3:00	*30*	20	20	*30*	30	20	20	20	4:00	20	20	20	*20*	40	20	20	*30*	5:00	6:00	20	20	15	*20*	40	20	20	20
1100	4:00	*20*	20	20	*20*	30	20	20	*30*	5:00	20	20	15	20	40	20	20	20	6:00	7:00	20*	20	12	20	*40*	20	20	20
1200	5:00	20	20	15	20	*40*	*20*	20	*30*	6:00	20	20	12	20	*40*	20	20	20	7:00	8:00	15	20	12	15	*40*	20	20	20
1300	6:00	20	20	12	20	*40*	*20*	20	*30*	7:00	20	20	12	15	*40*	*20*	20	20	8:00	9:00	15	20	12	12	*40*	*20*	20	20
1400	7:00	20	15	12	15	*40*	*20*	*20*	20	8:00	20	15	12	12	*40*	*20*	*30*	*20*	9:00	10:00	20*	20*	10	12	*40*	*20*	*30*	*20*
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1700	10:00	20	15	10	12	20	20	*30*	*30*	11:00	20	15	10	10	20	*20*	*30*	*30*	12:00	1:00	20	15	10	10	20	20	*30*	*30*
1000	11:00	*20*	15	12	10	20	15	*30*	20	12:00	*30*	15	12	10	20	12	*30*	20	1:00	2:00	20	15	12	10	20	15	*30*	20
1900	12:00	*30*	15	12	10	20	12	20	20	1:00	*30*	20	12	10	20	12	20	20	2:00	3:00	*30*	20*	12	20	20	12	15	20
2000	1:00	*20*	20	15	10	15	12	12	20	2:00	*20*	20	15	10	20	12	12	20	3:00	4:00	*30*	20	15	10	20	10	12	20
2100	2:00	20	20	15	10	15	10	10	20	3:00	20	20	15	10	15	10	10	20	4:00	5:00	*30*	20	15	10	20	10	10	20
2200	3:00	20	20	20	10	15	10	10	20	4:00	20	20	20	10	15	10	10	15	5:00	6:00	20	20	20	10	20	10	10	15
2300	4:00	20	20	20	10	15	10	10	15	5:00	20	20	20	10	15	10	10	15	6:00	7:00	20	20	20	10	20	10	10	15
MAY		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN			ASIA FAR EAST	EUROPE	S. AFRICA	CARIBBEAN S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.

*Look at next higher band for possible openings.



Heath HD-1481 remote coax switch

Hams have long known that Heathkit has produced some of the most memorable and valuable pieces of Amateur Radio gear, from transmitters and receivers to simple but useful station accessories.

A few years ago Heath came out with their first remote coaxial switch, the HD-1480. The beauty of this switch was that it allowed you to remotely select five antennas from a single feedline. This represented a significant cost savings in that you'd have to run only one coax line out of the shack to the switch; usually antenna feedlines from this point are much less. It also improved reliability in that instead of dealing with as many as 10 PL-259s (five antenna feedlines at two each) coming out of your shack, you'd have only two connectors.

The HD-1480 used a wafer switch and a stepping motor. Two problems, however, limited the switch's usefulness. The first was that you couldn't remote the switch much more than 200 feet from the shack due to power supply limitations. (Heath did beef up the supply to correct this in all but the longest runs.) The second problem was that the installation required the use of a separate control cable to run the switch, making installation more difficult and adding to the overall expense — control cable isn't cheap.

The new HD-1481 remote coax switch solves these two problems in an interesting and innovative way. The control cable has been eliminated by using the coax itself to carry the switching instructions — but more on how this is done later. You can remote the switch almost any distance away from the control unit because there's very little voltage drop in the heavy conductors used in coaxial cable. Installation is greatly simplified and consequently less expensive.

description

The remote coax switch comes in two units. There's the switch that you remote at the antenna site and the power supply/control unit that goes in your shack. The only interconnection necessary is the coaxial cable.

The remote unit uses two heavy-duty 24-volt DPDT relays as a switch. The unit is weather protected by a high strength black plastic cover, and all openings are sealed with silicone to ensure weatherproofing. It mounts via a single U-clamp arrangement.

The power supply/control unit includes a 24-volt power supply and a rotary switch that selects the appropriate polarity voltage to make the switch work.

theory of operation

I'm sure that by now you've either figured out how you can select four antennas with this switch or that you're fully confused. Fear not. Here's how it's done:

In switch position 1, no voltage is applied to the coax and Antenna 1 is selected. In position 2, the 24-volt AC is rectified to produce a positive 24-volt DC voltage. This is injected into the coax and energizes Relay 2 at the remote switch end and Antenna 2 is selected. In position 3, this is reversed: negative 24 volts is injected into the coax, energizing Relay 1 and routing the RF to an unenergized Relay 2 and Antenna 3. Finally, Antenna 4 is selected by sending AC to two diodes in the remote switch that energizes both relays. Sounds simple, doesn't it?

construction

Typical of a Heathkit, the instruction manual is clearly written, explained, and fully illustrated to minimize confusion and troubles during construction. I have the impression that Heath has gone to considerable time and expense making the manual easier to use than the ones that accompanied its kits of 10 to 15 years ago. (Maybe this is a sad commentary on the level of technical awareness of the average Ham nowadays?) The circuit diagram along with the exploded drawing should be more than enough to allow completion of this project with a minimum of fuss and trouble.

After completing a couple of tests you're ready to install the remote switch, a few antennas, and presto — that's it!

use

I've installed the switch out at my 160-vertical, which is located over 200 feet from the house. For starters, I've installed 80 and 40-meter antennas and have had no problems using either low or high power switching between any of the three antennas.

Heath includes self-adhesive labels so you can mark the antenna by both band and design, a neat extra touch.

I expect that this switch will give the average user thousands of hours of reliable use with few, if any, problems.

For more information, contact Heathkit, Benton Harbor, Michigan 49022.

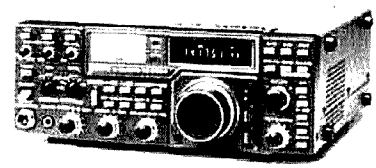
—NIACH



new IC-751A base station transceiver

The ICOM IC-751A 100-watt HF base station trasceiver and general-coverage (100 KHz-30 MHz) receiver incorporates the high performance features of the IC-751 with new, improved features requested by hams worldwide. This newly-designed, top-of-the-line HF transceiver includes the following:

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Optional filters include the FL-52A CW 455 KHz at 500 Hz, FL-53A CW N 455 KHz at 250 Hz, FL-63A CW N 9.0106 MHz at 250 Hz, FL-33 AM 9.010 MHz at 600 Hz, and CR-64 high-stability 30.72 MHz crystal filters.

The IC-751A will be available in April 1986 and will be displayed in the ICOM booths at the Dayton Hamfest (April 25-27).

For details, contact ICOM America, Inc., 2380 116th Avenue NE, Bellevue, Washington 98004.

Circle #317 on Reader Service Card.

current probe

The new MFJ-206 Antenna Current Probe determines the current distribution and RF radiation pattern of antennas, transmission lines, ground leads, building wiring, guy wires, enclosures, shields, etc. It monitors RF currents by sensing the magnetic field around a current

carrying conductor. It uses an electrostatically shielded ferrite core tuned circuit, and FET RF amplifier, and an operational amplifier meter circuit for excellent sensitivity and selectivity.

It can be used to adjust an antenna for maximum efficiency, gain, and F/B ratio to improve



DX, and to determine whether a ground system is effective so you can radiate more power. You can also use it to determine the best place to mount a mobile antenna on a vehicle for a stronger signal and eliminate RFI by pinpointing leaky shielding. It can even be used as a sensitive tuned field strength meter.

The MFJ Antenna Current Probe is powered by a 9-volt battery and covers 1.8 to 30 MHz in five ranges. It includes a telescoping antenna for the field strength meter, and sensitivity and tune controls are provided. Also included are an on/off switch, a power LED indicator, and an internal meter zero adjust.

The retail price of the MFJ-206 is $79.95.

For information, contact MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, Mississippi 39762.

Circle #316 on Reader Service Card.

antenna traps

G2DYM Aerials, makers of the well-known Anti-TVI trap dipoles, has announced the availability of new traps for 10 and 15 meters. A 6-inch length of aluminum tubing at the end of each trap facilitates the construction of two- and three-element rotary tribanders, rotary tribander dipoles, and trap verticals, either quarter or half-wave in height.

Prices are as follows: 10 and 15-meter traps, $16 plus $5 shipping; kit of four traps for rotary dipole, $54 plus $8 shipping; kit of eight traps for 2-element tribander, $105 plus $10 shipping; kit of twelve traps for three-element tribander; and kit of two traps for vertical, $27 plus $6 shipping.

For information, contact G2DYM Aerials, Uplowman, Tiverton, Devon, England EX16 70H.

Circle #315 on Reader Service Card.

buying, installing marine electronics

In his new book, *The Straightshooter's Guide to Marine Electronics, Motor Boating & Sailing Magazine* columnist Gordon West speaks frankly buying and installing marine electronics gear.

According to West, "More and more mariners are doing their own installation of marine electronics. In my book I tell them where to get the best deal, how to buy the equipment, and finally, how to install it so it meets the criteria to be covered under a manufacturer's warranty. I also talk about the necessary FCC licensing as well as steps to avoid electrolysis."

Topics covered include depth sounders, handheld and 25-watt VHF radios, marine SSB radios and ham radios; VHF and MF direction fiders, LORAN equipment, satellite navigational receivers, and automatic pilots, among others.

Personally autographed copies are available from Gordon West Radio School, 2414 College Drive, Costa Mesa, California 92626 for $9.95 plus $3.00 postage and handling.

Circle #314 on Reader Service Card.

cable and connector guide

Nemal Electronics international has published a comprehensive guide for the selection of electronic wire, cable, and connector products. The 32-page guide contains detailed specifications and illustrations of over 550 items, along with cable construction and performance charts, and a complete tooling cross reference.

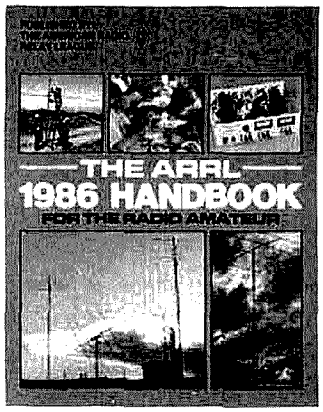
Among the 32 product categories listed are fiber optic cables and connectors, plenum cables, satellite control cables, and numerous RF and data connector types. Nemal's new cable and connector selection guide is available for $4.00 (credited with a $50.00 order).

For information, contact Nemal Electronics, 12240 NE 14th Avenue, North Miami, Florida 33161.

Circle #313 on Reader Service Card.

replacement mic for discreet communication

Ace Communications, Inc. has introduced the IECS-200 Inter-Ear Communication System. The IECS-200, which replaces any HT's speaker microphone, allows the user to speak as well as listen through the earpiece, protecting the privacy of communications



STATE OF THE ART

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TM-201B, TR2G/3600A, TM2570A/50A/30A
TW-4000A Kenwood Service/Repair
TH21/31/41AT, TM-211A/411A & TS-711A/811A

YAESU
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FT-726R, FRG-9600, FT-270RH, FT-2700RH
YAESU FT-1/2/703R, FT-270/RH, FT-1903/1123
ICOM IC2AT, IC02AT, IC-04AT, IC-A2/U16
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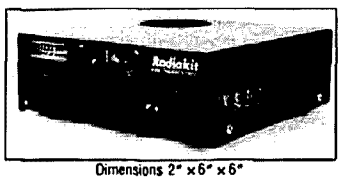
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For information, contact ACE Communications, Inc., 22511 Aspan Street, Lake Forest, California 92630-6321.

Circle #312 on Reader Service Card.

146 and 440 MHz mobile antenna

Austin's new 19-inch Model 500C antenna is designed using state-of-the-art technology currently being used in the cellular radio field. Ruggedly built with a low design profile, it takes advantage of several patent-applied-for techniques that enhance its radiation efficiency. The antenna uses the standard Motorola vehicle mount (not supplied).

On 146 MHz, it uses the vehicle body as a ground plane and is a 1/4 wave vertical with an elevated feedpoint. This technique brings the main lobe down to 16 degrees above the horizon; this lower angle of radiation improves the antenna's ability to get into distant repeaters. The standard 1/4 wave vertical has a main radiated lobe of 60 degrees above the horizon (the 5/8 wave is 22 degrees). Overall bandwidth is around 20 MHz; the antenna is rated at 100 watts.

On UHF, the 500C operates independently of the vehicle and is a 1/2 wave element. Tuning of the 1/2 wave stainless steel whip is achieved in the re-entrant cavity.

The retail price is $49.95; the Motorola MAG Mount is priced at $39.95.

For information, contact Austin Custom Antennas, P.O. Box 357, Sandown, NH 03873.

Circle #311 on Reader Service Card.

new Yaesu duplexer

Yaesu Electronics has announced the release of the new AD-2 Duplexer for the FT-2700RH Dual-Band FM Transceiver and FT-726R VHF/UHF All-Mode Transceiver. The AD-2 provides for semi- or even full-duplex VHF/UHF cross-band operation with a single 2-meter/70 centimeter dual-band antenna. The single antenna serves for both transmitting (on one band) and receiving (on the other band), *simultaneously*. Band-to-band isolation of more than 50 dB assured minimum receiver interference between bands. At high power (up to 50 watts), there is minimal insertion loss of transmitted power or received sensitivity.

specifications:

pass bands:	140-150 MHz and 400-450 MHz
maximum power:	50 watts
insertion loss:	VHF less than 0.3 dB; UHF less than 0.5 dB
impedance:	50 ohms
VSWR:	less than 1.2:1
receive isolation:	50 dB

For details, contact Yaesu Electronics Corporation, 17210 Edwards Road, Cerritos, CA 90701.

Circle #310 on Reader Service Card.

newsletter index

After serving the owners of Yaesu equipment for 14 years, Milt Lowens, N4ML, has announced the termination of the publication of the *FT Newsletter*, the official journal of the International Fox-Tango Club, which he organized in January 1972. All of the back-issues of the Newsletter have been republished in calendar-year volumes, mostly in booklet form; each has its own index.

To simplify the task of selecting the volumes most appropriate to individual needs, Fox Tango has also published a comprehensive 32-page cumulative index covering the years 1976 through 1985 in detail, and summarizing the years 1972 through 1975. Most articles are grouped according to model number (FT-101, FT-757, etc.); within such groupings, newsletter articles are listed chronologically by year and page; by topic (such as user report, modifications, etc.); and by title and author's call sign.

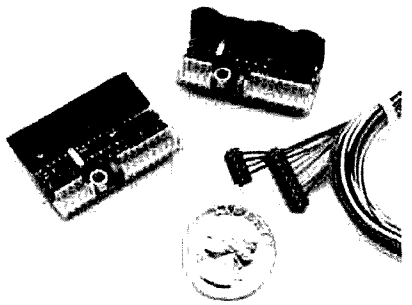
The price of the index (including a rebate certificate creditable towards the purchase of newsletter volumes) is $4.00 postpaid in the United States and Canada (elsewhere, $5.00).

For information, contact Fox Tango Corporation, Box 15944, West Palm Beach, Florida 33411.

super-small encoder-decoder for portables

Communications Specialists, Inc., of Orange, California has announced availability of their TS-32HB Super-Microminiature Programmable Encoder-Decoder for handhelds. The TS-32HB comes in two different configurations to take advantage of the limited space available in the newer super-small handheld radios. The TS-32HBH measures 1.5 x 0.65 x 0.65 inches. The TS-32HBL, lower in profile, measures just 1.5 x 1.2 x 0.4 inches.

Programming the 32 available CTCSS EIA tones is done through a five-position *DIP* switch mounted on the board, and installation is simplified by the use of two plugs with color-coded cables attached. A crystal-controlled clock oscillator allows excellent stability under all conditions, and sensitivity is rated at 6 mV RMS for use with the lowest output receivers. Decode bandwidth is ± 1.5 Hz. The TS-32HB provides an adjustable sinewave output and encodes with an accuracy of ± 0.1 Hz maximum at 40



degrees C to + 85 degrees C. Output level is 6v P-P across 10k.

Priced at $64.95, the TS-32HB is in stock for immediate delivery and is covered by a full one-year warranty. A catalog is available on request.

For details, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665-4296.

Circle #307 on Reader Service Card.

Commodore/ICOM interface

Microcomputer Electronics Corporation has announced the release of the new MEC 71$_{IC}$, a computer control interface for Commodore 64/128 computers and ICOM R71A and 751. The MEC 71$_{IC}$ is easy to use; screen menus guide the user through operation of the system. Features include UTC time display, frequency display/control with 10 Hz resolution, mode display/control including narrow filter status, and single-page viewing of ICOM's 32 internal memories. Full control of the ICOM's 32 internal memories, frequency stepping with selectable steps, and complete VFO and memory control, including VFO/memory exchange, are included.

The MEC 71$_{IC}$ is designed to be used with the ICOM EX-309 module, available at low cost from any ICOM dealer. The ICOM EX-309 is easy to install in the radio. The MEC 71$_{IC}$ plugs into the expansion port of the 64 or 128 computer and a cable exiting from it plugs into connector provided with the ICOM EX-309 module.

The MEC 71$_{IC}$ includes a 90-day warranty and a comphrehensive user's guide for detailed information on system operation.

The MEC 71$_{IC}$, priced at $199, is designed and manufactured by Microcomputer Electronics Corporation and is distributed by the Electronic Equipment Bank, 516 Mill Street, Vienna, Virginia 22180.

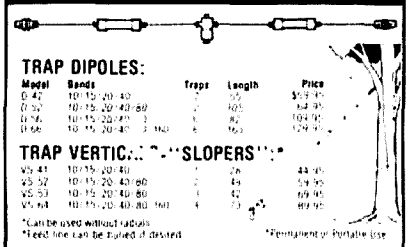
Circle #308 on Reader Service Card.

infra-red sensing digital thermometers.

North American SOAR has announced the release of four infra-red sensing digital thermometers.

Model TX-700L is a general-purpose, handheld, battery-operated portable instrument that can measure an object's temperature with a 6-inch (150mm) spot size at a distance of 16 feet (5 meters). Model TX-700S uses an LED as a spot marker, making aiming easy and sure.

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Model	Bands	Traps	Length	Price
D-42	10-15-20-40	2	66'	$59.95
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D-53	10-15-20-40-80	3	132'	104.95
D-66	10-15-20-40-80-160	6	263'	129.95

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VS-41	10-15-20-40	1	28'	49.95
VS-52	10-15-20-40-80	2	44'	59.95
VS-53	10-15-20-40-80	3	62'	69.95
VS-64	10-15-20-40-80-160	4	73'	89.95

*Can be used without radials
**Feed line can be buried if desired

*Permanent or Portable use

ALL TRAP ANTENNAS are Ready to use - Factory assembled - Commercial Quality - Handle full power. Comes complete with Deluxe Traps, Deluxe center connector, 14 ga Stranded CopperWeld ant. wire and End Insulators. Automatic Band Switching. Tuner usually never required. For all Transmitters, Receivers & Transceivers. For all class amateurs. One feedline works all bands. Instructions included. 10 day money back guarantee!

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Model	Band	Length	Price
D-10	10	16'	19.95
D-12	12	21'	19.95
D-15	15	33'	24.95
D-20	20	66'	24.95
D-40	40	132'	34.95
D-80	80	264'	34.95

Includes assembly instructions. Deluxe center connector, 14 ga Stranded CopperWeld Antenna wire and End insulators.

COAX CABLE: (Includes PL-259 connector on each end)

Type	Length	With antenna purchase	Separately
RG-58	50'	$8.95	$11.95
RG-8	50'	12.95	16.95

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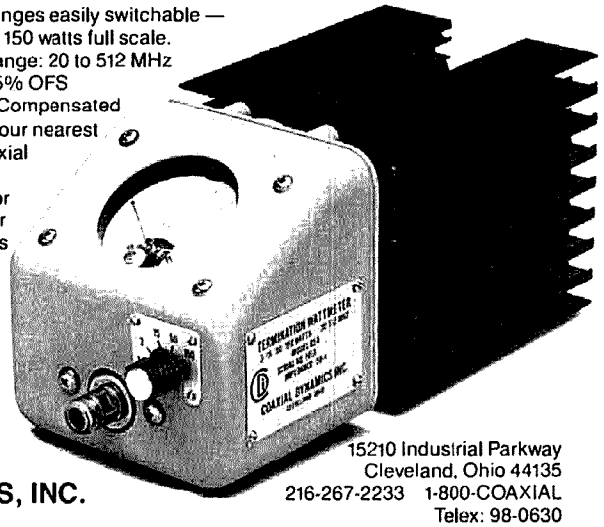
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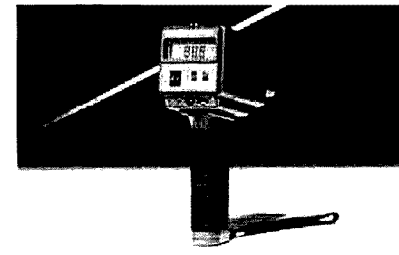
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NEW products



Models TX-710L and TX-710S are monitor versions (i.e., no pistol grip) of the TX-700.

The TX-700 Series units are small in size, light in weight, and easy to use. These units measure an object's temperature without touching and are highly accurate with resolution to 0.1 degree C. A Data Hold function is trigger activated and readings are viewed on a large 3-1/2 digit LCD in approximately half a second. The TX-700 Series has a high-low limit set capability with alarm output; it also has an analog signal output. An automatic "low battery" indication appears in the LCD readout when the battery's voltage falls below operating level. The TX-700L/S can be AC operated using an adapter provided. The price of the TX-700L is $1470.00 and the TX-700S is $1495.00.

For more information contact North American Soar Corporation, 1126 Cornell Avenue, Cherry Hill, New Jersey 08002.

Circle #306 on Reader Service Card.

simplex autopatch and HF base station control system

The Com-Shack 64 has been designed to give the ham shack a new dimension in user control. The full-featured simplex auto-patch and HF remote base operate under control of the Commodore 64 computer. A clear digitized human voice announces your call sign and alerts you to an incoming call. All parameters such as timing windows and time-out controls are adjustable from a user-friendly menu. A real-time menu displays all system parameters. Call waiting and last number memory features are included. The autopatch works on any phone line in either tone or pulse mode. A Yaesu 757 and a VHF/UHF transceiver are all that are required to complete the setup.

A fast-scan and a slow-scan mode provide remote tuning on all bands. The hardware interface board plugs into the I/O port of the Commodore 64 or 128. All hardware and cables and connectors are supplied along with a program

DX Alpha Delta Model DX-A 160-80-40 METER QUARTER WAVE TWIN SLOPER ANTENNA

The Model DX-A combines the tremendous firepower of the quarter wave sloper with the wide bandwidth of a half wave dipole. Simple to install, quick to tune. Proven longhaul DX performance.

- Installs like an inverted-V dipole. One leg for 80 meters (67') and the other leg for 160/40 meters (55'). Fed with a single 50 ohm coax. SO-239 connector provided on mounting bracket.
- Configuration provides wide bandwidth on all three bands. Typically 70 kHz on 160 meters, 200 kHz on 80 meters and full band on 40 meters. Much wider than most other loaded slopers, dipoles or verticals. Tuner usually not required.
- Model DX-A also operates on 30-17-12 meters. VSWR of less than 2.5:1. Easily matched with a tuner.
- High-power operation. Rated at 1500 watts P.E.P. output. No traps to break down. A single "ISO-RES" isolator-resonator is used in the 160/40 meter leg.
- Current lobe up high for maximum radiation and excellent DX performance. Can be installed from 25 to 40' high.
- The Model DX-A Antenna is fully assembled, uses all stainless steel hardware, a UV-protected "ISO-RES" coil, #12 copper wire and is rated for severe environments. Specially coated wire disappears from your neighbors' view.

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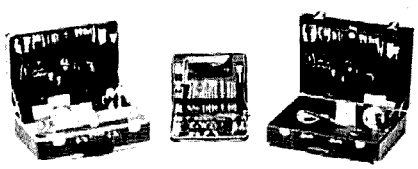
disk and instruction manual. Easy to install, the system is priced at $229.95

The Com-shack 64 is available from Engineering Consulting, 583 Candlewood St., Brea, California 92621.

affordable tool kits

Jensen Tools, Inc. has introduced a new economical line of electronic/electrical tool kits for students, hobbyists, and in-house service departments. While not as comprehensive as the professional Jensen line, the new Telvac kits include tools for most service needs and come in a choice of attache or zipper-style cases.

The tool selection includes more than 30 standard tools for servicing electronic and electrical equipment as well as for model building. Tools are supplied in single-pallet attache-style cases of wood/vinyl or aluminum/vinyl construction, or in a heavy-duty zipper pouch case.



Telvac Kits are ideal for those seeking a highly functional and attractive kit at a minimum investment. For more information, contact Jensen Tools Inc., 7815 S. 46th St., Phoenix, Arizona 85044.

Circle #304 on Reader Service Card.

new commercial license materials

In response to popular demand, Ameco Publishing has rewritten two Commercial Radio Operator Question and Answer License Guides for two popular commercial exams: the General Radiotelephone Operator's License and the Marine Radio Operator's Permit.

For preparing for the General Radiotelephone exam, *The Commercial Radio Operator Q&A License Guide for Element 3* (Cat. No. 9-01, $5.95) contains over 270 questions taken from the latest FCC syllabus. Detailed answers, written in Ameco's proven, easy-to-understand style, are included.

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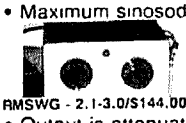
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113

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114

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NEW products

examination, plus additional charts and handy reference material.

Ameco also recommends its Commercial Radio Operator Theory Course (Cat. No. 15-01, $8.95). More than 450 pages long, it contains 21 lessons and 600 FCC-type multiple choice questions. The book is a complete course in commercial radio; no previous technical background is required.

For details, contact Ameco Publishing Corp., 220 East Jericho Tpke., Mineola, New York 11501.

Circle #303 on Reader Service Card.

Yaesu newsletter

International Radio, Inc., publishers of the ICOM and Kenwood *Newsletters* for the last five years, has now added the *Yaesu Owners' Newsletter*, formerly a Fox Tango publication, to its list of monthly publications. Founded by Milt and Ida Lowens approximately 13 years ago, *The Yaesu Newsletter*, like its companion publications, functions as a worldwide owner's information exchange. Each of the three newsletters features information gathered from readers, manufacturers or their agents, the IRI Service Laboratory, and Amateur Radio magazines around the world.

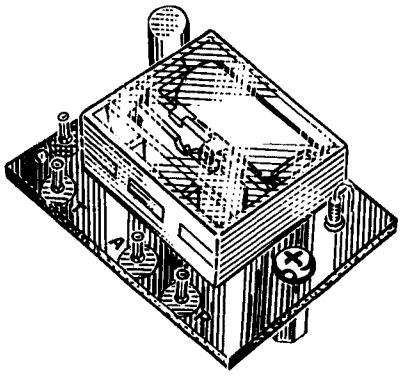
Back Issues of the Fox Tango *Yaesu Newsletter* are available from 1972; a cumulative index covering the years 1972 through 1985 is also available. Back Issues of the IRI *ICOM Newsletter* and *Kenwood Newsletter* are available from 1980; cumulative indices are also available.

For information on these publications, send an SASE to International Radio, Inc., 1532 S.E. Village Green Drive, Suite L, Port St. Lucie, Florida 33452.

Circle #301 on Reader Service Card.

new T/R relay module

Hamtronics, Inc. has recently developed a low-loss transfer relay module for use with 50-ohm coaxial cables at frequencies up to 1000 MHz. The special shielded relay, with gold-plated contacts for high reliability and long life, is mounted on a stripline PC board with solder terminals. The relay module, which measures only 1-1/8 x 1-5/8 inches, is easy to mount and connect. In a typical application, it would be wired to adjacent modules with miniature coax. To interface with larger cables, it's usually wired to appropriate connectors on the rear panel of the cabinet. Handy for T/R switching of devices such as the Hamtronics transmitter and receiver



modules, band switching of transmit and receive converters, selection of antennas, switching test equipment paths, etc., the TR Relay Module is rated to switch RF up to 25 Watts with a VSWR of only 1.3:1. Special rounded contacts and shielding between ports provides an isolation of 55 dB at VHF and 48 dB at UHF. The maximum usable frequency is 1000 MHz. Transfer time is a fast 15 mSec, which makes it useful in applications such as packet radio or simplex autopatch. Operating power is 13.6 VDC at only 40 mA, and on-board relay coil transient suppression is provided. Standoffs are provided for mounting the relay module.

The price of the TR Relay Module is only $24 in kit form and $34 assembled.

For information on the TR Relay Module and a 40-page catalog, contact Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535. (Enclose $1 to receive your catalog by return first class mailing. For overseas mailing, please send $2.)

200-watt linear amp

Microwave Modules Ltd., of Liverpool, England, has announced the availability of a combination 200-watt linear amplifier, the MML 144/200-S, with selectable input levels and a GaAsFET receive preamplifier for 144-MHz operation. This design provides an output power level of 200 watts that is fully compatible with 3-watt 144-MHz handheld transceivers or multimode transceivers having either 10 or 25 watts output. The input power level is manually switch-selectable to suit the transceiver in use, making this product appropriate for use with mobile, portable, or base station equipment. RF VOX and a manual PTT are provided.

The receive preamplifier uses one of the latest dual-gate GaAs FETs in a noise-matched configuration. Conversion gain is deliberately set at only 12 dB in order to achieve strong signal handling capability while maintaining an overall noise figure of below 1.5 dB. The unit is housed in a durable extruded aluminum enclosure measuring 13-3/16 x 6-15/16 x 3-13/16.

The MML 144/200-S is available from the "PX" Shack, Belle Mead, New Jersey 08502 and other Microwave Module dealers.

Circle #320 on Reader Service Card.

COMPUTER SMYTH
THE HARDWARE JOURNAL

Now there's a brand new magazine for the technically sophisticated microcomputer enthusiast who likes to build, customize and explore micro hardware at the chip and board level.

The publishers have recruited an outstanding staff of senior editors with more than twenty years of engineering, software and diagnostic experience in the microcomputer field. ***Computer Smyth*** is produced quarterly in Peterborough, New Hampshire, home of ten other microcomputing publications and a total of more than twenty internationally circulated publications.

Computer Smyth's primary interest is hands-on construction, modification and expansion of micros. We see the IBM PC phenomenon as a giant magnet or vacuum, dragging hardware and software talent into a vortex of activity that ignores and overshadows the line of new CPUs and peripheral hardware enhancements that are becoming available. We believe 32-bit architecture is the proper and exciting growth direction for micros and too little talent is being invested in that opportunity.

Computer Smyth's editors are also determined to cover all opportunities including the rich offerings of the IBM lines as they appear and especially to evaluate the so-called clones.

We believe magazines are hard-copy networks—or extensions of the central nervous systems of those who read them and interact with each other through them. The inter-stimulus factor accelerates each participant's learning curve, produces new combinations of ideas and new answers, and defines fresh problems. We are content and idea centered—not just a sales medium for consumer goods.

Who reads *Computer Smyth?* We're looking for the intelligent, technically curious and adventurous computer buff who isn't afraid to take the back off the case, who likes new experiences and digs into any device, unsatisfied until all its mystery is dispelled and its potential is fully in hand. Our reader is a craftsman who enjoys building, even while finding the adventure just a little scary.

Our first-year line-up included: The SC84 computer; a brand new Z80 system with exceptionally powerful peripheral possibilities and a plain English description of each and every capability of the machine and its operating system; an X/Y charter/plotter you can build for under $60 that will teach you a lot about how these devices work; a neat, powered wire-wrap tool for two hours of your time and a little more than the price of the tool's bit; an RGB color to composite converter board; Ed Scott's three part series on his 68000 computer; Ken Barbier on printer interfaces; a multipen plotter upgrade; a Data Destroyer (for super clean disks) a review of the DTC XT clone; an RMN converter; and an audio module for the Atari 8000.

Coming up in 1986: a switched power supply; an EPROM programmer; a safe, active circuit tracing tool; a silicon disk; how to identify and find microchips; how to buy a superplus keyboard; and builder reports on four single-board computers: Z80, 64B180, PC and XT.

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flea market

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CABLE TV CONVERTERS & EQUIPMENT: Plans and parts. Build or buy. SASE for information. C & D Electronics, PO Box 1402, Dept. HR, Hope, AR 71801.

WANTED: Anything made by Kinemetrics True Time, Inc. Receivers, WWV clocks, time code readers, manuals, incomplete assemblies, etc. C.K. Barber, 3112 Ursula Street, Aurora, CO 80011.

IBM-PC RTTY/CW. CompRtty II is the complete RTTY/CW program for the IBM-PC and compatibles. Virtually any speed ASCII, BAUDOT, CW. Text entry via built-in screen editor! 10,000 character transmit/receive buffers. Adjustable split screen display. Instant mode/speed change. Hardcopy, diskcopy, break-in buffer, select calling, text file transfer, customizable full screen logging, 24 programmable 1000 character messages. Now with WRU (simple mailbox). Ideal for MARS and traffic handling. Requires 128k PC, XT, AT, PCjr, PC-DOS, serial port, RS-232C TU! $65. Send call letters (including MARS) with order. David A. Rice, KC2HO, 7373 Jessica Drive, North Syracuse, NY 13212.

THE GOOD SAM HAMS invite RV operators to check in the Good Sam Ham net 14.240 *Sundays 1900Z also 3.880 Tuesdays at 2359Z. Net control N5BDN, Clarksville, IN*

OLD RADIO transcription discs wanted. Any size, speed. W7FIZ, Box 724 HR, Redmond, WA 98073-0724.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $10.00. Beginners RTTY Handbook $8.00. PO Box RY, Cardiff, CA 92007.

IMRA, International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, New York 10538.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

SELL RK4D32 Raytheon Transmitting tubes only $35.00 each. Inquire on others. Wanted Morrow BR 560 receiver with supply. Morrow SSB Transmitter, AC supply. M. Levy, 101 East Driftwood No. 26, Fredericksburg, TX 78624. Tel 512-997-2534.

WANTED: Western Electric tube manual. Also would like to contact groups specialized in home-brewing tube type audio equipment. Doug Yumoto, AG8L, 8258 Niles Center Road, Skokie, IL 60077. (312) 982-9683 (home) (312) 764-9186.

DISCOUNT CATV CONVERTERS/DECODERS and video accessories. Solar cells for remote power and battery charging applications. Send for free information and prices ... it could save you big money on your next purchase of these and other CATV items. Easy View, PO Box 221, Arlington Heights, Illinois 60006. (312) 952-8504. Ask for Rudy Valentine.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. Daily Electronics, PO Box 5029, Compton, CA 09224. (213) 774-1255.

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RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

BRAND NAME 5.25" COMPUTER DISKS. 10 disks in poly-pak: SSDD $7.99 DSDD-$8.99 includes sleeves, ID labels, tabs. Send check or money order to Micro-Applications, Inc., 915 E. Sheridan, Des Moines, IA 50316. Shipping add $1.50 for 1-5 items; $3.00 for 5 or more items. COD orders add $2.00. IA residents add 4% sales tax.

FAMOUS "FLORIDA SKIP" 2 meter repeater map. Now in its 5th printing. Fully up-to-date, with all the latest repeater listings and frequencies. Designed and edited by Bill Tucker, W4FXE. Available for just $.25 plus business size SASE to: FLORIDA SKIP, PO Box 501, Miami Springs, FL 33266

PRINTED CIRCUIT BOARDS and kits for QST articles. Call or write for information. A&A Engineering, 7970 Orchid Drive, Buena Park, CA 90620. (714) 521-4160.

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CHASSIS and cabinet kits. SASE K3IWK, 5120 Harmony Grove Road, Dover, PA 17315.

WANTED: Heath Curve Tracer IT-3121 kit/assembled. Mission School needs 10 units. Call Manila, P.I. 832-1773 or write Mission POB 1651 MCC Makati MM, P.I.

2 METER AMP KITS: 8877 legal limit kit $395. 3CX800A7 900W kit $325. Also HV power supplies, CX600N relays, parts and EME newsletter. SASE for catalog. 2 Meter EME Bulletin, 417 Staudaher St., Bozeman, MT 59715.

RIW-19 432 MHz beams by K3IPW. SASE for information.

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COMING EVENTS

Activities — "Places to go . . ."

CALIFORNIA: Flea Market/Boneyard Sale. Foothill College, Los Altos Hills. March-Sept. 2nd Saturday of every month 7 AM Sellers. 7:15 AM buyers. Talk in 145.27( ) or 147.570 simplex. FCC exams (408) 255-9000.

ILLINOIS: The Starved Rock Radio Club Hamfest, June 1, Princeton. Same place as last year. SASE please for complete registration materials, map, etc. SRRC, W9MKS, RFD 1, Box 171, Oglesby, IL 61348 (815) 667-4614

NEW HAMPSHIRE: Deerfield. The Hosstraders will present their Spring Tailgate Swapfest, Saturday, May 10 at Deerfield, NH Fairgrounds. Admission $2 per person; no extra charge for sellers. Friday night camping at nominalfee, but no entry before 4 PM Friday. Profits benefit Shriners' Hospital. Last year's gift: $11,754.46. Talk-in 146.40-147.00. New England's biggest Hamfest is sponsored by Bob, W1GWU, Joe, K1ROG, and Norm, WA1IVB. For map SASE to WA1VIB, RFD Box 57, West Baldwin, ME 04091.

RHODE ISLAND: The RI Amateur FM Repeater Service which operates the 146.16/76, 146.34/94 and 222.16/223.76 repeaters in RI, will hold their annual Spring Flea Market and Auction, Saturday, May 17, American Legion Fairmount Post 85, 870 River Street, Woonsocket. Flea market opens 9 AM and spaces are $5.00 each. Auction 12 Noon to 5 PM. Free admission. Food and beverages available. Talk in on 34/94 and 52 simplex. For information: Rick Fairweather, K1KYI, Box 591, Harrisville, RI 02830 (401) 568-3468 from 7-9 PM.

GEORGIA: The 1986 Atlanta Hamfestival, sponsored by the Atlanta Radio Club, July 19 and 20, Georgia World Congress Center. Over 150 exhibitors. Forums and a huge, air-conditioned flea market. For further information SASE to: Atlanta Hamfestival, POBox 77171, Atlanta, GA 30357.

ILLINOIS: The Western Illinois Amateur Radio Club will hold its 1986 Tri-State Swapfest in conjunction with a Flyin Breakfast sponsored by the Experimental Aircraft Association. May 18, Haerr Field, Taylor, Missouri (5 miles west of Quincy, IL). 9 AM to 3 PM. Tickets $1.00 each; 6/$5.00. Tailgate flea market spaces $2.00 advance, $3.00 at the gate. VEC exams offered for all license classes. Talk in on 147.03 repeater. For more information: Western Illinois ARC, PO Box 3132, Quincy, IL 62301.

NEW YORK: The 27th annual Southern Tier Amateur Radio Club's Hamfest, Saturday, May 3, Treadway Inn, Owego. Flea market opens 8 AM. Tech and non-tech talks. Vendor displays. Refreshments. 6:30 PM dinner by advance tickets only. Talk in on 22/82, 16/76 or 146.52 simplex. For further information SASE to PO Box 7082, Endicott, NY 13760.

MASSACHUSETTS: The Hampden County Radio Association will hold its annual Flea Market, Sunday, May 4, rain or shine, West Springfield Elks Lodge covered pavillion, Morgan Road, 9 AM to 3 PM. Admission $1.00. Tables $5.00 each. Dealer displays $5.00 per vehicle. Food and refreshments available. For more information contact Steve Nelson, WA1EYF (413) 596-8216.

INDIANA: The 40th annual WVARA Hamfest sponsored by the Wabash Valley ARA, Sunday, June 1, Vigo County Fairgrounds, Terre Haute. For more information SASE to WVARA, PO Box 81, Terre Haute, IN 47808.

NEW HAMPSHIRE: The 12th annual Eastern VHF/UHF Conference, May 16-18, Rivier College, Nashua. Friday night hospitality room. Saturday rap sessions. New this year ... a series of tutorial presentations for the newcomer to VHF/UHF. To preregister send $13.50 to David Knight, KA1DT, 15 Oakdale Avenue, Nashua, NH 03062 before May 5. Registration at the door $20.00. Saturday night banquet $14.00 payable by May 5. A special registration fee of $10.00 is available for the first-timer. Dormitory housing available at the college for a modest fee. Contact Lewis D. Collins, W1GXT, 10 Marshall Terrace, Wayland, MA 01778 (617) 358-2854 6-10 PM EST.

MASSACHUSETTS: The Northern Berkshire Amateur Radio Club's annual Spring Flea Market, Sunday, May 18, Dalton American Legion, Route 9, Dalton. Admission $1.00. Starts at sunrise. YLs and kids free.

ARIZONA: The Cochise Amateur Radio Association (CARA) will hold its 1986 Hamfest, May 2-4, Club's Training Facility, Moson Road and Rt. 90, Sierra Vista. No charge for tailgaters. Talk in on 146.52 and 146.16/76. Ham exams May 4. For more information: Don Morgan, W7ACI, (602) 458-5293 or CARA, PO Box 1855, Sierra Vista, AZ 85636.

MARYLAND: The Maryland FM Association's annual Hamfest, Sunday, May 25, Howard County Fairgrounds, West Friendship, 30 miles west of Baltimore. 8 AM to 4 PM. Inside tables $7.00 by advance registration only. At the door $10.00 if available. Donation $3.00. Talk in on 146.16/76, 222.16/223.76 or 449.1/444.1 MHz. For tables or information: Michael Cresap, W3IP, 1921 Pometacom Drive, Hanover, MD 21076. (301) 551-3567 6-10 PM.

COLORADO: The Pikes PeakRadio Amateur Association's Swapfest Saturday, May 17, 8 AM, Rustic Hills Mall, Palmer Park and Academy Blvd, Colorado Springs. Free admission. Table rentals $8.00 advance, $10.00 at the door. Talk in on 146.37/97. VE testing on site. For information or reservations call Al, N0CMW (303) 473-1660 or write PPRAA Swapfest, PO Box 16521, Colorado Springs, CO 80935.

CONNECTICUT: The Pioneer Valley Radio Association's annual Flea Market, May 4, 9 AM to 3 PM, East Hartford High School (formerly Penney High) 869 Forbes Street, East Hartford. Free parking. Dealer tables $8.00, setup 8 AM. General admission $2.00. Talk in on 146.19/79. Amateur radio exams given by the Newington Amateur Radio League VE team. Walk-

ins accepted. For information or table reservations: Dave Rose, KW1V, 13 Long Crossing Rd, East Hampton, CT 06424. (203) 267-8993.

OHIO: The Athens County ARA's 7th annual Hamfest, Sunday, May 18, 8 AM to 3 PM, City Recreation Center, East State Street. Admission $3.00 advance and $4.00 at the gate. This year the focus will be on the use of computer technology in ham radio. A special feature will be a live demonstration of packet radio. Those wishing to take license exams send completed form 610 and check for $4.25 payable to ARRL/VEC by April 15 to John Cornwell, NC8V, exam coordinator, 101 Coventry Lane, Athens, Ohio 45701. Advance registration requested but walk-ins will be accepted. Free flea market space for those bringing own tables. Talk in on club repeater 146.34/.94. For reserved flea market space: Sam Stewart, KA8NIE, 116 Franklin Avenue, Athens, Ohio 45701 (614) 592-5330. For further information or advance tickets write Carl J. Denbow, KA8JXG, 653 Morris Avenue, Athens, Ohio 45701.

OHIO: Portage Hamfair sponsored by the Portage ARC, Sunday, May 18, 7:30 AM to 4 PM, Randlph Fairgrounds, Randolph. Tickets $3.00 advance, $3.50 at the gate. Indoor/outdoor flea market spaces $2.00; with 8' table and one chair $6.00. For information or tickets (216) 274-8240.

KENTUCKY: The Northern Kentucky ARC's "Ham O-Rama '86", June 7 and 8, Best Western Vegas Convention Center, Erlanger (8 miles south of Cincinnati). Admission for both days $5.00. Children under 13 free. Open to public 8 AM. Indoor/outdoor flea market. Contact AF4Y or WD4PBF at gate for flea market spaces and prices. Major vendors contact Joe Dunnett, WA4WNF at (606) 371-2255 for additional information or write NKARC, PO Box 1062, Covington, KY 41012. Talk in on 147.855/.255 and 147.975/.375.

1986 "BLOSSOMLAND BLAST" Sunday, October 5, 1986. Write "BLAST", PO Box 175, St. Joseph, MI 49085.

PENNSYLVANIA: The Delaware County Amateur Radio Association is sponsoring their 7th annual Hamfest, June 8, Drexel Hill Middle School, State Road and Penn Avenue, Drexel Hill. Doors open 8 AM. Setups 7 AM. Admission $3.00. Indoor tables with electricity available by reservation $3.00 per space. Outdoor tailgating. Novice through Extra license exams starting 10 AM. Food and refreshments available. Talk in on W3UER, 147.96/36 C/L 224.5 MHz and 146.52 simplex. For registration/information: Hamfest, DCARA, PO Box 236, Springfield, PA 19064 or contact Barbara, N3DLG (215) 535-1616.

ILLINOIS: The Six Meter Club of Chicago is having its 29th annual Hamfest, Sunday June 8, Santa Fe Park, 91st and Wolf Road, Willow Springs, IL. (SW of downtown Chicago). Advance registration $2.00. At the gate $3.00. Swapper's Row. Gates open 6 AM. Plenty of parking, picnicking, displays in pavilion, refreshments. AFMARS meeting. Talk in K9ONA 146.52 or K9ONA/R 37-97. Advance tickets from Val Hellwid, K9ZWV, 3420 South 60th Court, Cicero, IL 60650.

NEW YORK: The Rochester Hamfest and Atlantic Division ARRL Convention, sponsored by the Rochester Amateur Radio Association, May 16, 17 and 18, Monroe County Fairgrounds, Rochester. Tickets $5.00 advance; $7.00 at the gate. Huge outdoor flea market $5.00 per space. Special indoor flea market for non-commercial exhibitors. For tickets: Hamfest Tickets, 174 Croydon Rd, Rochester, NY 14610. For information: Rochester Hamfest, 300 White Spruce Blvd, Rochester, NY 14623 or phone (716) 424-7136. Talk in on 146.28/88.

NEBRASKA: The Ak-Sar-Ben Amateur Radio Club Auction, May 18, Radial Social Hall, 1516 NW Radial Highway, Omaha. Equipment check-in 8 AM. Auction begins 9:15 AM. Talk in on 146.34/.94. For information: Greg Zimmerman, N0BTN (402)895-5219.

MICHIGAN: The Wexaukee Amateur Radio Association's 26th annual Swap Shop, Saturday, May 17, Wexford Civic Arena, jct. of N. Mitchell and 13th Street. 8 Am to 2:30 PM. Admission $2.50. Food and refreshments available. Talk in on WD8RZL 97/37. For more information: Wexaukee ARA, PO Box 163, Cadillac, MI 49601.

NORTH CAROLINA: The Durham FM Association will hold its annual Hamfest & Computerfest, Saturday, May 24, lower level of South Square Mall, Durham, 8 AM to 4 PM. Admission $4.00 at the gate. Covered flea market. Tables will be available. Free parking. Talk in on 147.825/225. For more information: D^{e}F^{a}M^{i}A^{s}, PO Box 8651, Durham, NC 27707 or Mick, W4ZUS at (919) 544-3556.

OHIO: The Sandusky Valley ARC's annual Hamfest, May 18, American Legion Home, 2000 Buckland Avenue, Fremont. Advance tickets $2.50; at the door $3.00. 8' table space $6.00. Free trunk space space, free parking. Talk in on 52 simplex or 146.31/91. For tickets or more info SASE to Pat D. Keating, WB8KWD, 615 Lime Street, Fremont, Ohio 43420.

NEW YORK: LIMARC sponsors ARRL outdoor Hamfest, Sunday, May 18, NY Institute of Technology, Northern Blvd, Rt. 25A, 1 mile east of Glen Cove Road, Old Westbury. General admission for hams $3.00. 9 AM to 3 PM. Sellers car space $5.00. 7:30 AM, no reservations. Call Hank, WB2ALW nights for additional info or LIMARC Infoline (516) 796-2366.

PENNSYLVANIA: The 32nd annual Breezeshooters Hamfest, Sunday, June 1, 9 AM to 5 PM, White Swan amusement Park, Rt. 60 near Greater Pittsburgh International Airport. Free admission and flea market. Family park. Registration $2.00 each. 3/$5.00, 7/$10.00. Covered vendors area by advance registration. Mobile talk in on 28/88 and 29.000 MHz. For more information: William Hall, Sr, K3VSL, 3103 Wainbell Avenue, Pittsburgh, PA 15216 (412) 531 4827.

CALIFORNIA: FCC exams. Novice-Extra. Sunnyvale VEC ARC. (408) 255-9000 24 hour. 73, Gordon, W6NLG, VEC

NORTH CAROLINA: Raleigh, the City of Oaks and the Raleigh Amateur Radio Society presents the 14th annual RARS Hamfest, NC State ARRL Convention and Computer Fair, Jim Graham Building, NC State Fairgrounds, Hillsborough Street. Advance registration $3.50 until April 7. $5.00 at the door. Flea market space, one table and 2 chairs (ours only) $6.00 each. FCC exams by pre registration prior to April 1. Contact John Johnson, WM4P direct. Free welcoming party in Graham Building Saturday night, 7 - 10 PM. Wouff Hong Ceremony. Talk in on 04/64 and 28/88. For more information: Rollin Ransom, NF4P, 2447 Fairway Drive, Raleigh, NC 27603. (919) 779-5021.

MASSACHUSETTS: The MIT UHF Repeater Association and the MIt Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday, May 21, 1986,7 PM, MIT Room 1 134, //Mass Ave, Cambridge, MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-5820/646-1641 or Craige Rodgers at 225-6616. Exam fee $4.00. Bring copy of current license, 2 forms of picture ID and completed form 610 (available from FCC in Boston, 223 6609).

OPERATING EVENTS

"Things to do . . ."

ARMED FORCES DAY: In recognition of the 37th anniversary of this event, Amateur Radio Station W4QDR, located North side aboard Naval Air Station Memphis, Millington, Tennessee, will be operated by sailors and Marines on Saturday, 17 May from 1400Z to 2200Z. For information on W4QDR, NAS Memphis contact: Station Custodian, Chief Petty Officer Bob Donan, KA4FAL, (901) 872-2007.

QRP ARCI Spring CW Contest, April 19 to April 20. For information Eugene Smith, KA5NLY, Chairman, POBox 55010, Little Rock, AR 72225.

UNION ELECTRIC Ham Radio Club will operate KA0AWS on May 18 from 1800-2300 UTC to honor all the employees of Union Electric Co, the utility serving parts of Missouri, Illinois and Iowa. Send contact number and large SASE (39 cents) for certificate to: KA0AWS Henry G. Schaper, Sr. 241 Tapestry Drive, St. Louis, MO 63129.

NATIONAL 6 MTR INVITATIONAL NET Activity Day Contest, 1400Z May31 to 2400Z June 1. Exchange call, SIN No. and grid squares. Send contest logs by July 1, 1986, to Lisa Lowell, KA0NNO, POB 249, Ft. Lupton, CO 80621

ST. CHARLES ARC will operate WB0HSI from the annual Lewis and Clark Days Festival, St. Charles, MO, 1300-2100Z, May 17-18. For certificate, send large SASE to St. Charles ARC, PO Box 1429, St. Charles, MO 63302

ARMADILLO COUNTY, TEXAS will become county number 3,077 from March 2 through December 31, 1986. Governor Mark White will proclaim that any Amateur Radio Operator operating along the Texas Independence Trail during that time frame may broadcast that s/he is in Armadillo County, Texas.

W.I.N.O., the Wireless Institute of Northern Ohio, an organization sponsored by the Lake County ARA, will commemorate Ohio Wine Month operating a special events station, KO8O, from a winery in Madison, Ohio, Saturday, June 7 and Sunday, June 8. A special certificate is available from KO8O WINO Weekend, 7126 Andover Drive, Mentor, Ohio 44060. Please send legal sized SASE.

ARMED FORCES DAY The US Naval Reserve Readiness Command will operate special event station W4NUS (Navy United States) from the battleship USS North Carolina at Wilmington, NC. May 17 1400Z to 0200Z. QSL to: USS North Carolina, Box 417, Wilmington, NC 28406. SASE not required for your special QSL.

OWENSBORO ARC will operate K4HY from 0000Z May 10 to 0530Z May 11 to celebrate their International BBQ Festival. 7245 phone. Certificate for SASE via N4EKG, 1615 East 23rd Street, Owensboro, KY 42301.

SCHOLARSHIP AWARD: The Atlanta Radio Club is pleased to announce its 1986 scholarship awards program. Two sums of $1250.00 each will be awarded to the winners. Applicants must be licensed Radio Amateurs graduating from high school and entering an accredited college or university as freshmen for the first time in 1986. Judging is based on school grades, citizenship, ham radio achievements and financial need. For application blanks write: Phil Latta, W4GTS, 259 Weatherstone Parkway, Marietta, GA 30067.

THE FOUNDATION FOR AMATEUR RADIO, INC., a non-profit organization with headquarters in Washington, DC, plans to award 21 scholarships for academic year 1986-87. Licensed Radio Amateurs may apply for these awards if they plan to pursue a full-time course of studies at an accredited university, college or technical school. For additional information and application form send letter prior to May 31, 1986 to FAR Scholarships, 6903 Rhode Island Avenue, College Park, MD 20740.

Derby and District Amateur Radio, incorporating Derby Wireless Club 1911, will be celebrating its 75th anniversary during 1986. The Society plans at least one event per month throughout the year each from a different location with the City of Derby. The callsign to listen for will he GB3ERD.

1986 marks the 50th anniversary of the Greater Cincinnati Amateur Radio Association. A number of special events are planned. Watch for announcements here.

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THE GUERRI REPORT

by Ernie Guerri, W6MGI

major advances in fiber optic systems —

Throughout the world intensive work is being done to tap the potential offered by fiber optic communications and data transmission systems. The progress has been rapid and impressive; in the United States and Japan, vast optical networks are taking shape that will handle much of the national communications load by the early 1990s.

The development of critical components has been key to this progress. The use of single-mode fibers has reduced transmission losses to the region of 2 dB/km. Single-mode fibers have very small core diameters, which means that the light doesn't reflect off the core walls and cause corresponding losses. Other key components include transmitting and receiving illumination and detection semiconductors, low-loss fittings and couplers, and integrated optical signal processors and data switches.

The bandwidths and data rates associated with fiber optic systems are dazzling, with several hundred MHz being common. These systems offer the further advantages of being small, lightweight, low in power radiation, EMP resistant, and very difficult to intercept or tap.

During the past several months, researchers at GTE have published work that shows just how fast this technology is progressing. The GTE investigators have fabricated a laser diode with a cavity only 0.2 square microns. This small geometry has demonstated switching rates of 20 Gigabits per second, and would theoretically allow data transmission rates 5 million times greater than a conventional phone line. This remarkable component was fabricated from Indium Phosphide (InP), a successor of sorts to GaAs.

As if all of this weren't enough, extensive work is being done to develop RF-to-optical converters, optical switches, and other devices aimed at eliminating the electronics altogether. Just when we thought that electronics was the way to go!

thanks for the memories

One form or another of electronic memory has found its way into nearly every type of consumer and industrial product: microwave ovens that store the cooking time for many foods in a permanent memory . . . scanners, Hi-Fi receivers, VHF-UHF transceivers and HTs, and so on — not to mention the proliferation of memory types and sizes available for computer applications.

Basic memory these days generally consists of CMOS structures organized in various binary configurations. The largest of these arrangements has typically been 256K x 1, used in advanced microcomputers like the IBM, Apple, and HP machines. These, in turn, are organized into even larger grouping — up to 1 Megabyte (1 million x 8 bits). Toshiba has recently announced the limited availability of its TC511000 series of 1 Mbit (1 million x 1) CMOS memories. Since most computers use a memory structure that calls for 9 bits (8 data bits plus 1 check bit), it will now be possible to have 1 Mb of random access memory using only 9 ICs. It has just recently become possible to break the so-called 640K computer memory barrier with bank-switched cards that allow well over 1 Mb of memory in microcomputers that nominally can address only 1 Mb of memory. The new Toshiba chip will help make possible IBM PC-type computers with processing power that is today associated with machines costing over $200,000.

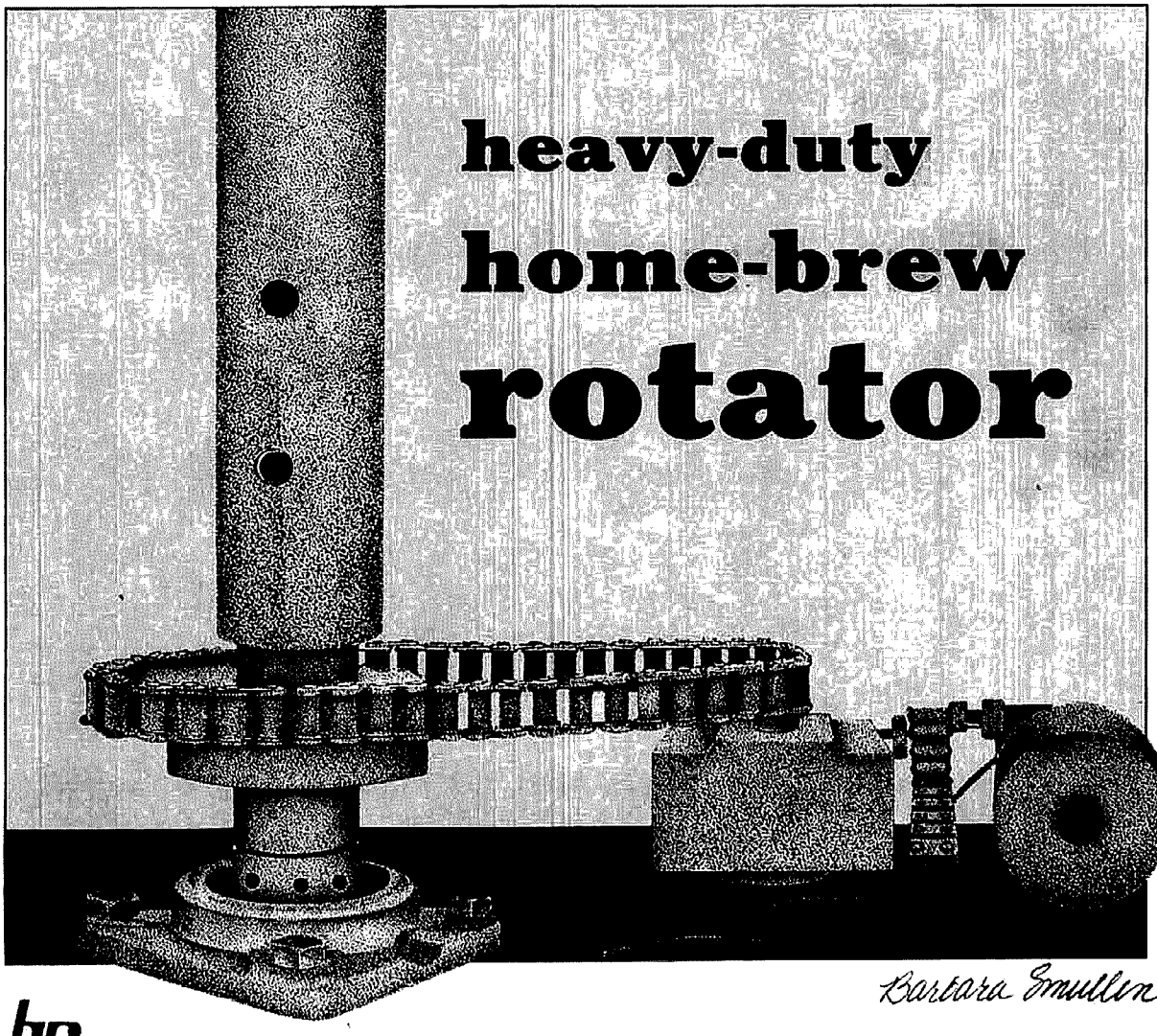
Other types of memory devices have made equally impressive strides. Floppy disks capable of storing over 6 Mb have been tested and are about to become available. Hard disks with 20 to 50 Mb are now common adjuncts for microcomputers, and industry sources say that within two years, 100 Mb drives will be "standard" on many advanced PCs. Plug-in expansion cards containing 20 Mb hard disks are now available at prices comparable to the cost of a single 360K floppy drive a few years ago. On the horizon are optical disks similar to those used in audio CD players; some are now available as permanent back-up for computer data. These devices can be written only once, but have over 100 Mb of capacity and are quite cost-effective for this application. Good progress is being made on much larger optical disks, with storage capacities up to 1 Gb, and the ability to read as well as write. Optical memories offer the prospect of being able to hold all the data that a single user may ever need, on a single disk!

ultra-small transistor

Scientists at Arizona State University report that they have fabricated an operating transistor with a gate length of only 150 atoms. This is less than 1/10th the size of the average microwave GaAs FET. The device was made using a special very narrow electron beam to form the etching mask. Since frequency response is closely related to the dimensions of the gate, it is expectd that devices with such small geometries will provide good gain and low noise well into the millimeter (above 30 GHz) region.

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ham radio magazine



heavy-duty home-brew rotator

Barbara Smullen

hr

focus
on
communications
technology

ham radio magazine

JUNE 1986

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T. H. Tenney, Jr., W1NLB
publisher

Rich Rosen, K2RR
editor-in-chief
and associate publisher

Dorothy Rosa, KA1LBO
assistant editor

Joseph J. Schroeder, W9JUV
Alfred Wilson, W6NIF
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publishing staff

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assistant publisher

Rally Dennis, KA1JWF
director of advertising sales

Dorothy Sargent, KA1ZK
advertising production manager

Susan Shorrock
circulation manager

Therese Bourgault
circulation

cover art:
Barbara Smullen

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REFLECTIONS

Who is Dorothy Rosa?

For the past three years anyone who's telephoned or corresponded with the *ham radio* editorial department has probably had the pleasure of hearing the pleasant, patient voice of or receiving a well-wrought response from Dorothy Rosa, our Assistant Editor. Magazines, of course, represent team effort, a necessity in the consistent publication of issue after issue of timely, accurate information. Dorothy, KA1LBO, is an important team member who's involved in every aspect of production from initial edit to final proof just prior to press run.

Dorothy came to us at an important time in our development, replacing Marty Hanft, KA1ZM, and stepped into a position that requires an extremely fast learning curve. Deadlines and schedules in general do not conveniently wait for someone to come aboard and get up to speed. They're planned, they arrive, and must be met. Then they rapidly become a thing of the past, only to be replaced by new ones.

Dorothy, in a sense, is not a newcomer to radio. Though recently licensed, she shared, at an early age, in the excitement of radio. Her father, an avid experimenter and SWL'er, had the normal collection of vacuum tubes, countless bins and small wooden drawers of parts, as well as innumerable carcasses of radios and various sorts of instrumentation. She remembers playing in the tube-lit glow of his repair shop, the smell of solder ever-present. It was, she says, the stuff of magic because even with his eighth-grade education, he could put all the pieces together and make a radio — and receive signals from thousands of miles away.

At the time, the unfortunate and mistaken notion that radio was for boys and reading was for girls was prevalent, so Dorothy became a voracious and compulsive reader, and as she likes to say, "a passable speller and moderately able writer." Though her major and degree were in English, she studied "a little bit of everything" in high school and college because "it was all so interesting." After college, she worked for a medical magazine in New York City, then moved on to a weekly newspaper and finally to dailies.

She spent several years as an advertising copywriter and photographer, trying to live what she calls "the simple life — which was anything but simple" in a log cabin in Vermont. She taught high school English while still finding time to tend the gardens. She traded all that in for the present position at *ham radio* and we are much the better for it.

Presently, while raising two sons and juggling multiple issues of the magazine at the same time, she's working on upgrading her license. She says she feels that, in some very small way, she's working off a debt here at HR . . . to her father for giving her the world through radio *and to* the hams who sent word of her first child's birth to Vietnam in 1969.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

ARRL'S "NOVICE ENHANCEMENT" PROPOSAL IS NOW AN FCC NPRM, released by the Commission just in time for the Dayton Hamvention. As relaased, the Notice of Proposed Rule Making follows the League's original proposal very closely (see Presstop, July and August, 1985).

In Brief, The 10-Meter Novice Band Would Become 28.1-28.5 MHz with CW, RTTY, AMTOR and packet on the bottom 200 kHz plus SSB from 28.3-28.5 MHz; the 200 W power limit (Novices and Techs only) would remain. Full privileges on the entire 220 MHz band plus 1246-1260 MHz are also proposed, though with a 25 W power limit on 220 and 5 W on 23 cm.

Though 220 MHz is Included In The NPRM, There's A Serious Question as to whether it will remain as part of the final result. That band has been the subject of attack from other services in the past, and is presently part of an as yet unfinished FCC study. The latest threat is a Petition for Rule Making from the Association of Radio Reading Services, which proposes using frequencies in the 220-225 MHz band for a new service for the blind.

Though Originally Hailed As A Means Of Attracting New Novices, as written the proposal looks more effective as a means for retaining existing Novices. However, that in itself would be a worthwhile result as the Novice dropout rate has consistently been the highest.

Comment due date for the Novice Enhancement NPRM, PR Docket 86-161, is July 16. Reply Comments will be due at the Commission on August 20.

THE FCC HAS REVERTED TO ITS ORIGINAL DEFINITION OF "EMERGENCY COMMUNICATIONS" as defined in Part 97.3(w). In an April 14 news release, the Commission said it reinstated the old definition in response to petitions for reconsideration filed by the ARRL and David Popkin, W2CC. The League's petition maintained that the new definition, which had been part of BC Docket 79-47, was too specific. The issue really came to a head during the Mexican earthquake emergency last year, when a number of non-Amateur organizations were heard using the Amateur bands for support and coordination activities. However, the issue of emergency communications is still far from resolved and will be addressed again soon.

ANY TWO REPEATERS WITH AN INTERFERENCE PROBLEM ARE "EQUALLY AND FULLY RESPONSIBLE" FOR RESOLVING IT, the FCC decided in its Report and Order on PR Docket 85-22. If only one of them is coordinated, however, the uncoordinated repeater has "primary responsibility" for resolving the problem. In the same action the Commission also dropped its restrictions on repeater power and antenna height. Effective date for the Report and Order is July 12.

RFI PROBLEMS HAVE BROUGHT THE SHUTDOWN AND FINING OF A CANADIAN AMATEUR, VE3SR. Ruling on a nuisance complaint brought by neighbors, the judge found that the neighbors' rights had been "unreasonably invaded by conduct which forms a basis for liability" and issued a permanent injunction against VE3SR operating a transmitter! In addition, he fined VE3SR $2500 for creating "inconvenience and interference with the enjoyment of various pieces of electronic equipment," $58.50 in Special Damages, plus costs and interest.

Canadian Amateurs Are Appalled At The Adverse Decision, as VE3SR had been supported by the Canadian Department of Communications as well as the CRRL, CARF, and a number of both Canadian and U.S. Amateur clubs. VE3SR had made a strong effort to resolve the problem, which involved a number of the neighbors' home entertainment devices, but met with little cooperation from them. At presstime he had not decided whether to appeal the decision, in light of the considerable investment he's already made in defending himself.

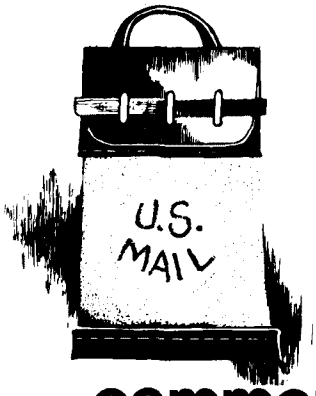
PRB-1's First "Victory," Lakeside Park, (Kentucky), Is Still Not Resolved. WM4T still has to appear before a Federal District Court judge to determine what constitutes a "reasonable" antenna, and now the city has adopted a new antenna ordinance that it calls "fair and reasonable" but area Amateurs feel is little better than the old one.

Even A 2-Meter Vertical Can Get An Amateur In Trouble with the law. KA7TVC put a 2-meter vertical on the roof of his Kirkland, Washington, apartment building with the owner's permission, then found himself in trouble with the city over a law requiring engineering drawings, a public hearing, and a $350 hearing examiner review fee before installation of any transmitting antenna! The law applies only to transmitting antennas, so the same antenna installed by a scanner user would not have been covered! KA7TVC is appealing his citation, and hopes with ARRL and PRB-1 help to be able to have the ordinance rescinded on the grounds that it is vague and unenforceable.

ROY NEAL, K6DUE, HAS BEEN NAMED "RADIO AMATEUR OF THE YEAR" by the Dayton Hamvention. The NBC newsman received the award for his many years of personal and public dedication to the Amateur Radio Service, and specifically for his key role in initiating Space Shuttle Amateur Radio activity. He's also been directly involved in many film and video promotions on behalf of Amateur Radio. U.S. Senator and former astronaut John Glenn paid a surprise visit to the Hamvention banquet to help honor Roy, who was also the banquet's keynote speaker.

The 1986 "Special Achievement Award" Went To Fr. Michael Mullen, C.M., WA2BQW, President of the International Mission Radio Association. He was cited for his dedication to IMRA and his accomplishments in supplying Amateur equipment to missionaries worldwide.

Doug DeMaw, W1FB, Received the 1986 "Technical Achievement Award" for his numerous contributions to Amateur Radio communications technology. Congratulations to all three!



comments

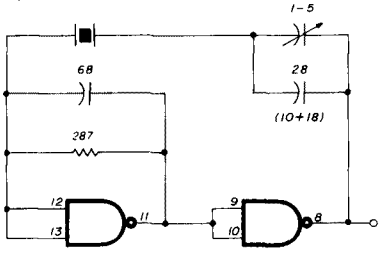
improving "a frequency and level standard"

Dear HR:

While I always find several articles of interest in each issue of *ham radio*, the January, 1986 issue — with PA0CX's article, "A Frequency and Level Standard" (page 10) — was especially timely. I have needed a frequency standard usable up to the 70-cm ham band for some time. To get one with calibrated amplitude was too much to miss, so out came my breadboard.

I've since made several improvements to the circuits that appeared in the article.

The first change was to use a different oscillator circuit. I couldn't get the author's circuit shown in **fig. 3** (page 12) to oscillate with any 1-MHz crystal that I owned. The circuit shown below has worked with all 1-MHz crystals



that I've tried. One advantage of this circuit is that it uses the 74LS00, which draws less current than the 7400. This aids frequency stability because the IC doesn't heat up as much. I measured approximately 15 Hz of drift at 200 MHz for a 7400, while the 74LS00 had less than 5 Hz from turn-on.

The second change is to substitute another 74LS00 for U3; this lowers current consumption and thus increases battery life.

I first built the circuit on a protoboard. The output amplitude was within 1 dB through 150 MHz, as the author, Hans Evers, had stated. This was without the use of any of the bypass capacitors shown in the schematic.

I then wired a final version on vector board for enclosure in a shielded box. This version did not have the flat amplitude response that protoboard had. In fact, there were a number of peaks in the response. Adding the 2.2 μF capacitors (especially the ones on U2 and U3) improved the response at the higher frequencies, but the problems at 100 kHz and 1 MHz mentioned by the author still remained. This was cured by placing a 2.2 μF capacitor from the 78L05 input to case ground (*not* circuit ground). I'm not sure if the regulator was oscillating or if there was a ground loop problem. I wired the IC grounds together with one long piece of No. 26 wire, although the protoboard has a good, wide low-impedance ground bus. The author's version appears to be wired in a similar fashion.

With the above changes the output amplitude was -73.0 ± 0.5 dBm from 100 kHz to 150 MHz and ± 1 dB to 220 MHz, as measured with an HP-71100A Spectrum Analyzer. For such a simple circuit, I couldn't ask for more.

The schematic (**fig. 3**) omits one important detail: pins 2, 3, 6, and 7 of U4 and U5 need to be grounded. This is correctly shown in **fig. 4** (page 14), but is easily overlooked when wiring up the circuit from the schematic diagram.

Thanks for the excellent article. I'll be waiting for the next issue of *ham radio*.

Steve Lund, WA8LLY
Santa Rosa, California 95401

on novice proposals

Dear HR:

I've been reading some negative comments about the proposals before the FCC to enhance Novice privileges. Both the ARRL's RM-5038 and my own RM-5025 request small band segments, restricted power, and the inclusion of digital modes. We never intended to "give the store away," nor do we expect the FCC to grant all the privileges we requested. Our common interest is the growth of our hobby in a responsible way.

I fail to see what those who oppose these proposals fear. There will be no loss of "higher class" privileges. Who protested the access we gained to 30 meters, or 12 meters, or 902 MHz? The Novice operators have gained nothing. The Technicians gained a band, but it is not really an incentive to Novices to upgrade. The big incentive to Novices seems to be 2-meter FM, and that is why my proposal did not include that band.

The ARRL's proposal includes more kHz than mine and is less practical with its inclusion of the 1246 MHz band. I have seen my proposal mentioned in only one publication (*Worldradio*, September, 1985) and have received no comments. My aim is to offer non-hams a Novice license that will attract intelligent adults and young people. We need computer enthusiasts, story tellers, teachers, students, accountants, laborers, etc., etc. Our record shows that we cannot lure them in any numbers with our CW-only Novice license.

Teach a Novice or an upgrade course like I do and you will see the work that goes into each new ham. You old-timers may forget that ham radio and electronics are a new language to the non-ham. We can help squash the myth that equipment is too expensive if we offer new hams phone and digital modes as well as CW. The equipment will be of more use and will [therefore] seem a better value. Besides, Novices really do *earn* a license.

If, after reading this, you still think enhanced Novice privileges are a bad idea, go read the actual proposals and then comment to the FCC.

Larry W. Garens, WD5H
Brady, Texas

turning that big array

Homebrew this inexpensive, quiet rotator

I recently bought, and was faced with the problem of rotating, some large five-element monobanders made by a manufacturer whose promotional materials emphasize their products' ruggedness and resistance to wind damage. Until that time, my largest antenna had been a loaded 2-element beam for 40 meters that could be handled by a HAM IV rotor. But my new antennas were much more massive — for example, my booms measured 3 inches (7.6 cm) in diameter, thus presenting a wind load that was not only large, but unbalanced as well because the boom length each side on the mounting mast differs slightly. This imbalance increases the stress on any rotator in high winds. Wise heads, then, agreed that disaster would be certain unless I obtained a heavier rotator.

Unfortunately, such rotators are expensive. In the past, Amateurs used World War II surplus prop-pitch motors designed for bomber aircraft; these, however, are no longer plentiful and also come with their own set of problems. I concluded that it should be entirely possible, in a city as diverse as Toronto, to find enough surplus mechanical components to homebrew a rotator at reasonable cost. What follows is not a detailed, step-by-step construction article, but instead a description of a general approach that can be adapted to suit the materials at hand.

mechanical design

A trip to the local surplus store uncovered a 20-to-1 worm gear reduction drive and a powerful single-phase, capacitor-start induction motor with integral gear reduction. The shaft speed of the motor is 56 rpm at 50 in.-lb (5.6 N.m) torque. It draws 2.1 amperes at 115 VAC. These two parts form the heart of the rotator and should be purchased first because the rest of the design will depend on how these have to be mounted and coupled together. The type of worm drive shown in fig. 1 is ideal, in that the input drive shaft is horizontal, and the output drive shaft is vertical; this facilitates running a chain drive to the mast. Furthermore, it mounts with four bolts at the bottom, making it easy to mount on a horizontal plate. Because a worm drive cannot be back-driven, no brake is required.

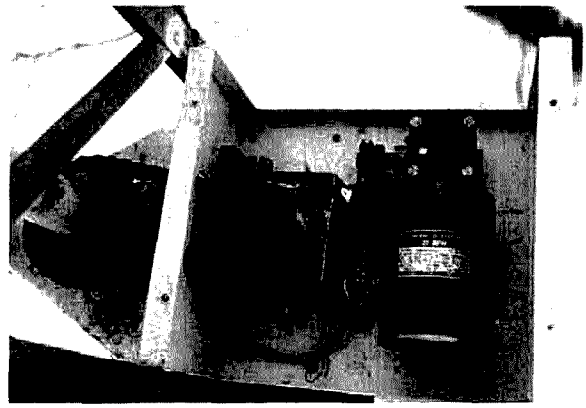


fig. 1. Worm gear reduction drive; input drive shaft is horizontal, output drive shaft is vertical.

The type of motor used, common on the surplus market, is available with various speed and torque ratings. The exact rating isn't too critical, as long as the product of speed and torque is at least 1500 and the

By Victor Mozarowski, VE3AIA, 1 Belgrove Drive, Islington, Ontario M9B 1S2, Canada

speed is low enough so that the rest of the drive train can reduce it to the desired antenna rotation speed.

drive motor and gearbox

A tremendous amount of torque is not required; the ruggedness of the gearbox is far more important. After all, if static friction is overcome, the antenna will eventually come up to speed, at which point its inertia will help keep it moving. Since most installations now use a high-quality ball-type mast bearing, starting would be a problem only if this bearing were iced up solid. Be that as it may, with the sprocket and gear ratios I used, and disregarding losses, the torque at the antenna mast works out to approximately 4500 in.-lb (508 N.m). (I suspect it might be possible to have a too-powerful motor that might damage the gearbox if the antenna mast were frozen.) There are formulas in machinery handbooks for calculating the size of components in the worm drive, but because these are based on a continuous running load and a certain working life, they're of little use if you want to know the ultimate yield strength under catastrophic load conditions. With carefully selected components, this type of failure would seem unlikely — and, in any case, as I'll show later, this type of failure isn't really catastrophic after all.

other components

As for the various other drive parts, suppliers sell a standard range of mechanical components, such as bearings, drives, gears, sprockets, and chains, at reasonable prices. The particulars of these components can be found in specialized catalogs (such as from Boston Gear Company, 14, Hayward Street, Boston, Massachusetts 01271), in the same way that we select standard values of electronic components from manufacturers' and distributors' catalogs. If a part isn't in stock, the supplier can almost always get it quickly through a distribution network. All the components used here except the motor and gearbox, which were bought at surplus, and the mast coupling, which was custom-machined, are standard components.

When I disassembled the gearbox, I found that its gears were also standard Boston Gear parts, so that in the unlikely event of a failure, it could also be repaired. This is one advantage of building your own rotator; because the parts are standard, reasonably priced, and readily available.

The mechanical parts required include the following:

- **Flanged cartridge.** A high-quality, sealed, ball-bearing unit set in a horizontal mounting flange (**fig. 2**), it can support both thrust and radial loads. The one I used could take about 2000 pounds (907 kg), although it's only a small unit and takes a 1-1/8 inch shaft (2.85 cm). Mounted at the bottom of the mast, it takes the full weight of the antenna array through the machined adapter shown in **figs. 2** and **3**. My unit's mounting holes were very close to the standard rotator mounting bolt pattern, so only a bit of

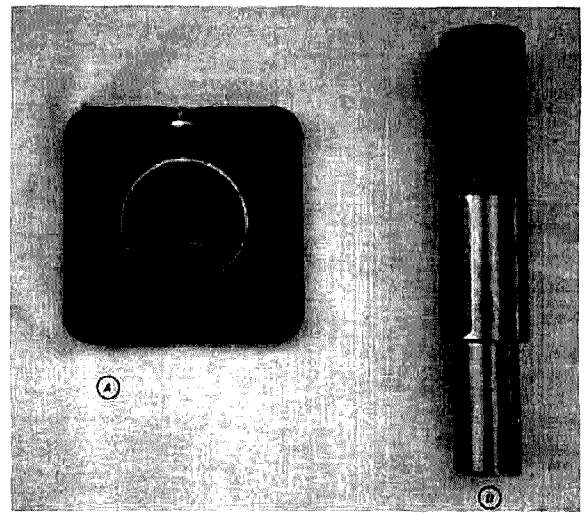


fig. 2. (A) Flanged cartridge is a high-quality ball-bearing unit set in a horizontal mounting flange. (B) Hollow steel pin secures mast adapter to the mast.

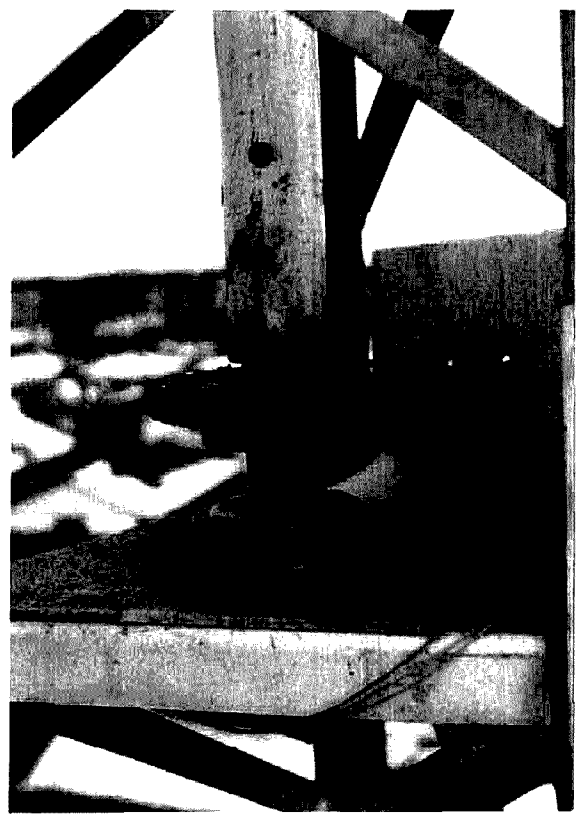


fig. 3. Completed mast assembly consists of chain drive steel pin and flanged cartridge.

filing was necessary on the tower's existing mounting holes.

- **Sprockets.** These are quite similar to bicycle sprockets. The type I used is available for various shaft sizes and is keyed for convenience in mounting and to prevent any slippage between shaft and sprocket. When the sprockets are pinned with a piece of key-stock, slippage is impossible; no clamping method can ensure this.
- **Rolled pin (or split dowel).** This is a hollow steel pin with a slot running the length of it. It holds the mast adapter (**fig. 2**) in the mast, and is driven with a sledge hammer into a hole drilled through mast and adapter. The slot allows it to compress as it's driven in, for a tight fit. There are two good reasons to use a rolled pin or split dowel rather than a bolt: one, there's absolutely no free play — and any free play in a large system is dangerous, because the antennas can slam back and forth, loosening or even shearing hardware — and two, there's no protrusion as long as the correct length is used. I used two 2-inch (5 cm) rolled pins, each measuring 3/8 inch (9.5 mm) in diameter, at right angles. One is visible in the lower hole in **fig. 3**; the hole above allows ventilation to protect the mast against condensation. The pins are available in various sizes and lengths (check the Yellow Pages under *Fasteners — Industrial*), but are probably not available at your local hardware store.
- **Gears for the direction potentiometer.** Small aluminum gears in various ratios that gear down the mast rotation to drive the direction-indicator potentiometer are available from hobby shops. Make sure both gears have the same tooth size (pitch). This is readily visible; inspect carefully.
- **Chain.** Chain is available in various sizes from the same sources as the sprockets (see above). For the final drive to the mast, I used a No. 50 chain, which also happens to be a common motorcycle chain. Note that it's much cheaper to buy this from a motorcycle parts counter than from a machinery supplier. It may also be possible to scrounge the short length actually needed from a motorcycle or bicycle repair shop.
- **Keystock.** This steel rod, square in cross section, is available in 1-foot (30.48 cm) lengths. You cut the length you need with a hacksaw. The keystock fits into a keyslot machined in the sprocket and the shaft to prevent slippage.
- **Mast adapter.** This is a straightforward piece of custom-machined steel (**fig. 2**). The end with the two transverse holes slides into the bottom of your mast and is secured with two rolled pins through mast and adapter. The length of this portion isn't too critical, though it probably shouldn't be shorter than 3 inches (7.6 cm), which is what I used, to minimize play. To determine the diameter for close fit, your machinist should measure the inside diameter of your mast, since it can vary significantly from nominal. Mine was 0.01 inch (0.5 mm) undersize, which means that if the adapter had been machined to nominal, it could not have been inserted! The sprocket fits into the middle part of the adapter, which should be long enough so you can line up this sprocket with the one on the gearbox. A keyslot should be machined to correspond to the keyslot in your sprocket. Be aware that sprockets for different shaft diameters will also use different key sizes. This portion was turned down to 1.5 inches (3.8 cm), which was the maximum shaft diameter for a keyed No. 50 sprocket with 24 teeth.

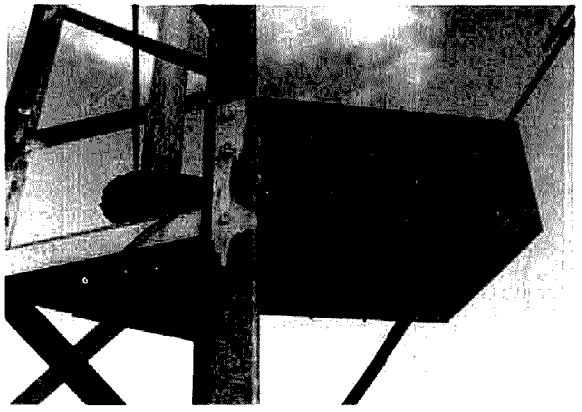


fig. 4. Bracing mounting plate to cross member prevents excessive flexing.

If you're using a large-diameter mast, I recommend machining a shoulder between the portion that slides into the mast and the portion that takes the sprocket. In this way, the shoulder takes the weight instead of the rolled pins or through-bolts.

The bottom part of the adapter, of smaller diameter, fits into the bearing in the flanged cartridge, resting on the shoulder formed by the middle part, which is of larger diameter. It should be long enough to protrude far enough below the rotor plate to mount a small gear for transmitting the motion to the direction pot. The end face was drilled and tapped in the center to take an insert onto which the gear could be mounted. Although the end could be machined down to the required 1/8 inch (3.1 mm) diameter, you would be unhappy if this little bit ever broke off. The insert is easy to machine.

The method described above for transmitting motion to the mast is superior to any mast clamping arrangement because the forces concentrated on this area are tremendous. The clamp on a popular commercial rotator has been known to loosen even in

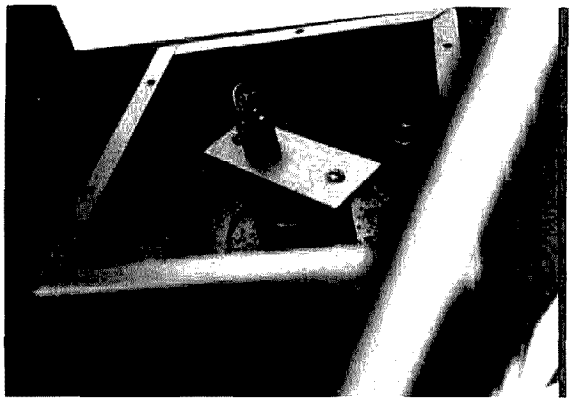


fig. 5. Rotation switch reverses output leads for reversing direction.

moderate winds, and the bolts have been known to shear altogether. A rotator manufacturer must accommodate a range of mast sizes, but you can make something that's perfect for the mast you're actually using.

As for having the machining done, any medium-size industrial town will have a number of shops that do general machining. I found one by driving through an industrial area and stopping at the first machine shop I saw; you may want to check the Yellow Pages or ask friends for their recommendations.

final assembly

When the adapter has been machined, take your mast to the shop and have the machinist drill the holes through both pieces together. This also allows you to check the fit.

My objective was a rotation speed of approximately 0.7 rpm , as a compromise between speed of rotation and starting-torque stresses on the tower. In the case of giant arrays, 0.5 rpm or less may be preferable.

The sprockets were selected this way: I wanted to use the heaviest chain to the mast for which sprockets were available that would fit both the 5/8 inch (1.58 cm) gearbox output shaft and the 1.5-inch (3.8 cm) mast adapter. The Boston Gear catalog showed that sprockets for No. 50 chain were the biggest that would allow this.

To minimize the stress on the gearbox, I wanted the maximum possible reduction between gearbox and mast. With standard keyed sprockets, this was 9 teeth to 24 teeth. A much lighter chain (No. 35) could be used between motor and gearbox. To achieve the desired rotation speed with the motor and gearbox I had, a ratio of 9 teeth to 15 teeth was used between motor and gearbox. Luckily, the 15-tooth sprocket was also the largest sprocket that could fit on the gearbox input shaft without scraping the mounting plate.

The rotor is assembled on a piece of 3/16 inch (4.76 mm) hardened aluminum plate bought from a junkyard (also a good source of gearboxes). If normal aluminum is used, I would recommend at least 1/4 inch (0.635 cm) thickness. Steel is preferable because of greater stiffness, but is much harder to work with, especially when filing or cutting, since it can't be cut on a shear. The mounting holes for the motor and gear drive must be slotted to allow the chains to be tensioned. Since the main drive chain is too heavy to tension directly by hand, a mechanical method had to be incorporated. **Fig. 1** shows the steel bar across the front face of the gearbox. Screws passing through it and the L-shaped blocks on either side of the gearbox are tightened to pull the gearbox in the direction away from the mast, thus tightening the chain. The gearbox mounting nuts and bolts are then tightened. The sprockets can be aligned by sliding them along the shafts, then securing them with the set screws. **Fig. 3** shows the mounting plate braced to a cross member across the tower face to prevent excessive flexing.

If you live in the North, make sure the grease in the gearbox doesn't solidify at low temperatures. Also, build or buy an enclosure to protect the rotator against rain. (I built another small box below the rotor plate to house the direction pot.) The mounting holes for the potentiometer bracket should be slotted to allow alignment of the gears, as shown in **fig. 4**.

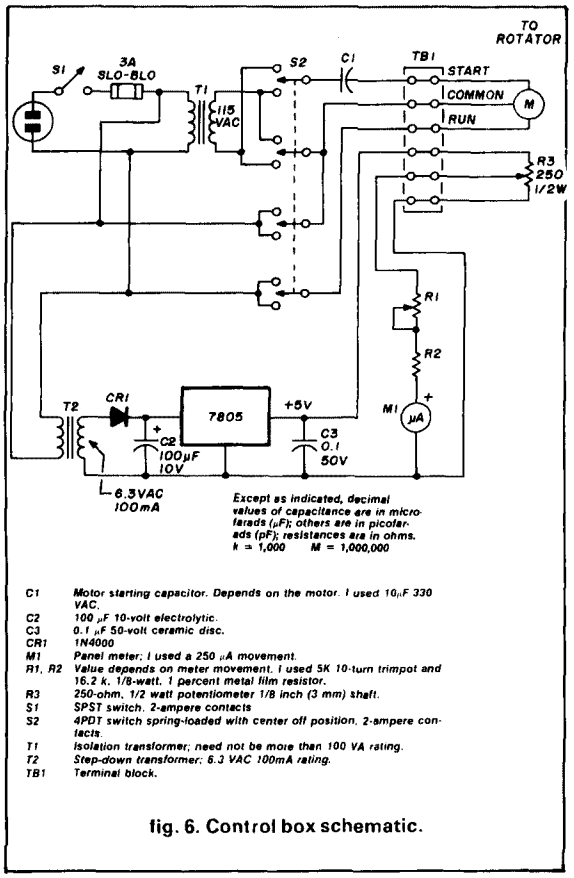
electrical design

For obvious reasons, the motor must be reversible. A capacitor-start motor can be reversed by reversing the current in the starting winding. To do this, I used a 1:1 isolation transformer in the starting winding; the rotation switch reverses the output leads for the reverse direction (**fig. 5**). The starting winding is the one with the higher resistance; the other is the running winding. Since the entire motor is rated at 2.1 amperes, I would think that a 100-watt transformer would be more than adequate. Usually only three wires are brought out from the motor, since a common return is used for both windings. If you're lucky and find a motor with the windings brought out separately, no transformer will be needed for reversing.

rotation switch and indicator pot

I didn't incorporate overtravel limit switches for a number of reasons. It was very awkward to mount microswitches and actuate them; also, many more heavy conductors would have been needed in the control cable. To minimize risk of overtravel, a spring-loaded rotation switch is used, with a center off position. Also, a meter was used for direction indication. Unlike a selsyn, this has no 0/360-degree ambiguity. If you want to use selsyns, install limit switches.

Since a pot turns only about 270 degrees, I used



C1 Motor starting capacitor. Depends on the motor. I used 10µF 330 VAC.
C2 100 µF 10-volt electrolytic.
C3 0.1 µF 50-volt ceramic disc.
CR1 1N4000
M1 Panel meter; I used a 250 µA movement.
R1, R2 Value depends on meter movement. I used 5K 10-turn trimpot and 16.2 k. 1/8-watt. 1 percent metal film resistor.
R3 250-ohm, 1/2 watt potentiometer 1/8 inch (3 mm) shaft.
S1 SPST switch. 2-ampere contacts
S2 4PDT switch spring-loaded with center off position. 2-ampere contacts.
T1 Isolation transformer; need not be more than 100 VA rating.
T2 Step-down transformer; 6.3 VAC 100mA rating.
TB1 Terminal block.

fig. 6. Control box schematic.

gears to couple to the mast. The exact gear ratio isn't critical, since full-scale deflection can be adjusted by the resistors in series with the meter (**fig. 6**). I recommended at least a ratio of 1:1.5, which provides for some travel past 0 and 360 degrees. (270 degrees × 1.5 = 405 degrees). Use a small pot with a 1/8 inch (0.318 cm) shaft, so standard Japanese gears can be used.

antenna direction

Antenna direction is displayed on a large panel meter. I painted over the original calibration, retaining only the scale, and used dry transfer lettering to re-label it in degrees. I wouldn't recommend using the original zero on the meter scale as your 0-degree azimuth point, but rather the next major scale marking. This is because a pot wiper seldom goes to 0 ohms, and you would risk damaging the pot or losing your calibration by having the pot rotate on its mounting if you accidentally went a bit past 0 degrees. I used a meter with a 250 microampere movement, but other movements (up to 1 milliampere) would be appropriate if the resistors in the control box were changed.

calibration

The direction indication can be calibrated as follows; it's best done on a sunny day. Take the control box outside for convenience. After deciding which scale markings on the meter will be your 0 and 360-degree points, turn the antenna for 0 degrees indicated. At this time the antenna heading isn't important because we're first calibrating the meter only for full-scale deflection. Place a long stick on the ground, in line with the antenna boom, or in line with the shadow of the boom. Rotate the antenna 360 degrees so the boom once again lines up with the stick. The antenna should have turned in a clockwise direction. Adjust trimpot R1 for 360 degrees indication on the meter. Double check the 0-degree point and repeat the above if necessary. Now you're ready to calibrate the true heading of the antenna.

Using one of the usual methods, rotate the antenna until it points to true North. The meter should show between 0 and 360 degrees. Loosen the mounting nut on the pot and turn the body until the meter reads 0 degrees. Your control box is now calibrated.

One of the nice features of this rotator is quiet operation. The tower doesn't resonate with the clang of a wedge brake disengaging and engaging, since no brake is provided or needed. Rotation is totally inaudible from the ground. Maintenance consists of keeping the chain greased and tight (though not *too* tight) and greasing the bearing once in a while.

I hope you have as much pleasure homebrewing and using your own rotator as I did.

ham radio

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secrets of successful low-band operation: part 2

Careful planning, quality components give DXing's "big guns" the competitive edge

This concluding section examines site requirements, lightning precautions, construction, maintenance, performance, and propagation. In addition, some of the most competitive stations reveal their plans for further improvements.

siting

Though most stations surveyed opt for the highest, quietest location possible, several work within the *same constraints* that the *majority of us* (the "little pistols") must deal with. Those surveyed were asked to describe their sites in terms of near and far field topography, obstructions, and noise.

Near field topography is a description of the contours of land in the immediate vicinity (i.e., within approximately one wavelength) of the antenna system. Of those who responded to this question, 83 percent described their near field topography as being flat or having a negative slope; the remaining 17 percent indicated a positive slope. It's important to remember that especially on the lower bands, optimum launch angles often exceed 30 degrees, so perhaps that mountain range in your backyard doesn't have as deleterious an effect as it might on the higher HF bands.

Far field topography in which the land rises at some distance should have an even less pronounced effect on the reception and transmission of low-band signals. However, the actual slope is important (in both the near and far field) and can be determined through the use of United States Geological Survey 7.5-minute (1:24,000) or 15-minute (1:62,500) maps.* Those surveyed described their far field conditions in basically the same terms as near field: flat or negative horizon. For example, they used the words "flat to ocean," "flat many miles," "mountains 26 miles away," "top of hill," "top of ridge," "flat and drops off," "flat to within 1/2 mile in all directions," and "80 meters above local terrain." So you see what most of the big guns have in common: a good site — topographically speaking, at least.

Obstructions. In terms of obstructions on those sites, the worst offender appears to be trees. Yet many of the low-band installations depend on these trees for supporting wire antennas. To date I haven't seen any definitive studies indicating whether the presence or absence of trees greatly affects patterns or performance of low-band antennas. (Any reader who has information on this subject, please contact me.)

Many of those surveyed indicated "none" when asked about obstructions; it's no accident that the FCC at its various monitoring facilities has extremely stringent requirements with regard to the height and location of any obstructions (even in what's defined as the far field). Apparently development of many of the high-performance stations involved site examination and evaluation as a first necessary step. An example of a good near and far field site is seen from the boom's perspective of W6NLZ's 80-meter Yagi (**fig. 1**).

Noise. The oft-spoken adage, "If you can't hear them, you can't work them," certainly applies to low-band operation. (There are some who have a particular knack for working stations that just aren't there, but that's another story). The limiting factor in the reception of signals is noise, be it man-made or atmospheric.

*Maps for areas east of the Mississippi are available in the United States from the United States Geological Survey, 1200 South Eads Street, Arlington, Virginia 22202. Maps for areas west of the Mississippi may be ordered from the United States Geological Survey, Box 25286, Federal Center, Denver, Colorado 80225. USGS maps are also sold by more than 1650 commercial dealers listed in the pamphlet, "Index to Topographic Maps," available without charge from either USGS distribution center listed above.

By Rich Rosen, K2RR, Editor-in-Chief, *ham radio*

After reviewing all the responses and assigning a range of 10 (no manmade noise whatsoever) to 1 (terrible), the average value appears to be around 8. Although most of the high-performance stations have a quiet noise environment, one of the top stations surveyed — with over 300 countries worked on 80 meters and well over 100 on 160 — described his location as "high noise," with a 115-kv high-voltage line within 0.5 mile. Don't give up hope for successful low-band operation if you have a high noise level. If it's the result of faulty electrical equipment, power companies can be helpful in tracking down the source of the noise and correcting the problem.

Other sources of interference can sometimes be located using portable equipment (for example, a pocket AM radio with a ferrite loop or circuit designed especially for that purpose). Noise cancellation techniques have been used for many years with some degree of success.[1,2,3] Many who use directional arrays (phased verticals, for example) probably notice that switching the antennas around produces a noticeable increase or reduction in noise level. Sometimes it's better to point the antenna's null at the interference, trading off a dB or two of gain from the preferred signal direction. A suggestion from some of the veteran 160-meter operators is for newcomers to be aware of noise from TV horizontal oscillators, either your own or your neighbors'.

lightning

The expression, "Into each life a little rain must fall" could be particularly applicable to big guns, whose large antenna systems are more exposed to the elements than others. Lightning poses a real threat to some of these stations, with their high towers and negative horizons. Precautions can be taken, however, and the danger lessened.

Two locations worthy of attention are in the shack and out at the antenna. Precautions taken at the shack include the simple expediency of disconnecting all transmission lines, grounding the equipment and/or antennas or a combination of both. Outdoor precautions include using antennas that are permanently DC grounded by design; for example, the base of a shunt-excited vertical is at DC ground. One method of DC grounding a horizontal wire antenna would be to attach a quarter-wave shorted stub (or odd multiple of a quarter wave) across the dipole's feedpoint and ground it at the shorting bar at the opposite end. This technique, primarily used to drain off static charge, works well only if the antenna is used over a narrow band of frequencies. Otherwise the short reflects back to the feedpoint as other than an open circuit and must be accounted for (in matching). Some of those surveyed were quite satisfied that the grounding system designed as part of their antenna installation would prevent any lightning damage.

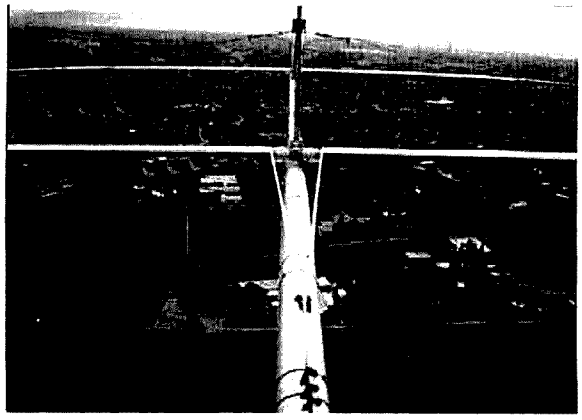


fig. 1. Sighting along boom of a KLM 4 element Yagi reveals unobstructed near field topography.

Even if a lightning discharge should occur at your antenna, shack damage is not inevitable. One "big gun" responding to the survey said he felt that by locating the antenna a considerable distance from the shack, the danger of damage occurring at the station end was greatly lessened.

Those who have taken direct strikes reported losses ranging from only a fuse box on one hand to almost total destruction of their home, with walls and windows blown out and ceilings collapsed. Anyone who's ever experienced even a very close lightning strike knows that this is a subject to be taken seriously, with every reasonable precaution applied.

construction

It's one thing to design, on paper, a three-element 80- meter Yagi. It's quite another matter to build, install, and keep it up. Wire antennas, though easier to install, still require an investment in quality materials, labor, and time.

Several of those surveyed built their own rotary Yagis. Noel, VE2HQ, described his efforts to construct his three-element, 515-pound, 43 square foot (windswept area) Yagi, the result of three years of planning, determination, and hard work, marked by several setbacks.

Working all winter in his basement, he built the linearly loaded elements, starting with 3-inch O.D. tubing and down to 0.5 inch in increments of 0.25 inch (11 different diameters). The linear loading was added 125 inches from the butt of each element by cutting the 2-inch diameter section, inserting a fiberglass tube and winding the coil with copper-clad wire. Additional element support was provided by double element guying on each side with Phillystran.**

The boom consisted of two 35-foot long flag poles

**Phillystran is a registered trademark of Philadelphia Resins Corp.

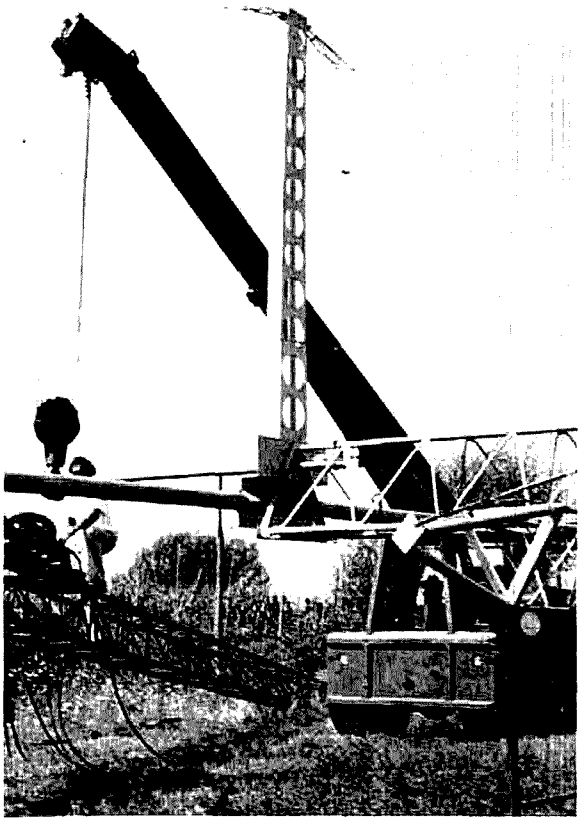


fig. 2. Vertical member and stays prevent element droop.

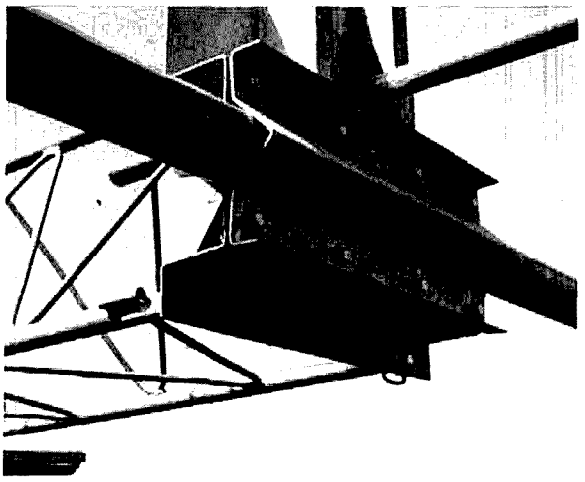


fig. 3. Method of attaching full length elements to boom involves use of 8 bolted through plate for mechanical rigidity.

(4.5 inch O.D., 0.25-inch wall) resulting in an overall length of 65 feet. At this point Noel had to consider what type of tower could support this heavy antenna. He purchased a commercial 60-foot tower designed for supporting large microwave antennas and fabricated a 50-foot triangular extension measuring 33 inches on a side to be mounted on top of that.

The first time Noel installed his antenna at 110 feet, it resonated too high in frequency and had to be taken down. It was made to resonate at 3625 kHz at 15 feet above ground so that it would resonate at 3785 kHz at its final height.

Chris, I5NPH, erected an even larger three-element Yagi using full-size elements. A careful examination of **fig. 2** provides an appreciation of the size of the components used in this antenna. The structure at the right that looks like a tower is the end of the boom; the large tubing at the left, which looks like a large boom, is one of the parasitic elements. Notice the size of the vertical strut used to support the long elements. **Figure 3** shows the method used to attach the parasitic elements to the boom. When one considers the difficulty of installing a completed antenna, eight bolts in the element-to-boom clamp don't represent "overkill."

While W6RJ put his three-element KLM Yagi together on the tower, K3ZO completed his Yagi's construction on the ground and raised it with a 164-foot crane. WA1EKV used ropes strung from his tower in a vee to slide and pull his Yagi to the top.

Because wire antennas tend to break at the worst times, a quick means of lowering and raising them is necessary. Ropes, pulleys, and continuous halyards are successfully used by some to provide this capability. In addition, if you want your antenna to stay up, use the best quality material. One of those surveyed has had success with white/blue marine rope. Polypropylene rope, though less expensive, disintegrates after exposure to sunlight. It might take several years, but it *will* deteriorate; be sure to wear gloves when working with old polypropylene.

installation and maintenance

Two more practical questions were asked of those questioned in this survey: "What precautions did you take in installing your system to make sure it would stay up?" and "What periodic maintenance procedures do you follow?"

In general the consensus of opinion can be summed up in a single word: *quality*. Though topnotch material costs more initially, it delivers years of consistent good performance and pays for itself.

The "big guns" offered specific advice on the subject of non-wire antennas (rotaries and towers). Here's what they recommend:

- Tighten all guy wires and clips to specification.
- Choose all materials carefully — anything that isn't aluminum or stainless steel will rust.
- Paint all towers and masts.
- Lubricate all moving parts properly.

With regard to wire antennas, they advise the following:

- Use quality halyards — for example, white/blue marine rope.
- Use steel line for the halyard section that goes over tree branches.
- Avoid the use of galvanized steel wire for radials.
- Protect all critical connections with a good weather-resistant sealant such as Coax-seal.®

Periodic maintenance procedures include:

- Tightening of guy wires, clips and hardware.
- Lowering wire antennas with pulleys to check the condition of the halyards, wire, and insulators.
- Cleaning insulators with a rag and water.
- Clearing foliage away from verticals.
- Repairing thin wire ground radial systems.
- Protecting systems with the same paint used on the undersides of boats.

Many of the big guns found that they experimented so frequently with new antennas that the short life cycle of their old antennas precluded the need for periodic maintenance.

performance

The next question posed was, "How does your present antenna system perform in comparison to others — either your own or the competition's?" In general most felt that the latest was the best. Specifically:

K3ZO found that his three-element KLM rotary at 140 feet was 25 dB better than a half-wave sloper and up to 8 dB better than his previously used three-element, 80-foot boom delta loop beam.

OH1RY found his full-size three-element Yagi outperformed a dipole by 20 to 30 dB on the long path shot to the United States.

4X4NJ uses four tilted verticals on 160 meters, fed in various phase combinations with passive reflectors to create an electronically rotatable vertical beam system. He feels "it's almost as good as having a full rosette of Beverages, with the added advantage of having a good low-angle rotatable transmitting antenna."

W2JB, another 160-meter operator, stressed the importance of being able to *hear*. "Most of the DX runs low power and signals are usually at the noise level. That's where you need your Beverage antennas," he wrote.

SP3GEM, who trades off between a vertical and a delta loop, doesn't see much difference between the two in the favored delta loop direction. However, he prefers the vertical in other directions — that is, where the delta loop exhibits a null.

KG7D, like many others, believes slopers provide good performance by launching a low angle signal, showing nulls off the sides, and having reasonable feedpoint impedances and consequently high efficiency.

One of the truly big signals from Japan on 80 meters, **JF1IST**, uses the Create Manufacturing CY-703 three-element Yagi and is very pleased with its electrical and mechanical performance. The specification sheet lists its forward gain, F/B ratio and 2:1 VSWR bandwidth as approximately 8.5 dBi, 20 dB and 100 kHz, respectively. JF1IST has his centered on 3.80 MHz, right in the middle of the Japanese SSB window.

W1NH, formerly W1SWX, can often be heard pulling out the rarest of stations on 80 meters, thanks to his three slopers and inverted vee from a 100-foot skyhook and six 1000-foot long Beverages. To illustrate its performance, he enclosed copies of QSL cards from VS5MC, HS5ABD (both short and long path) and JT0DAQ — all very difficult shots from the East Coast on 80.

K5UR attributes his success on the low bands to a superior receiving capability (Beverages once again). Considering that he holds the *CQ* Worldwide DX Contest U.S. record on 160 meters, his transmitted signal must also be outstanding to be heard above some of his competition on the East Coast. He does mention that his new 130-foot vertical appears to work noticeably better than his old 70-footer.

YU7PFR's three-element wire Yagi at 56 feet outperforms his previous single vertical or dipole while providing a 2:1 VSWR bandwidth of 250 kHz. With one of the strongest signals from Yugoslavia, he's consistently heard in the United States.

SM4CAN has noticed as much as a 6 dB improvement in signal level using his Bobtail curtain as compared to his gamma fed tower with 120 radials. However, on long path to the United States his delta loop performs better than the Bobtail *during the month of November.*

According to **SM6EHY**, it's possible to have too low an angle radiator. He says that his four-element phased vertical array just doesn't have that extra punch when high angle signals are propagated, and goes on to say that "the greatest advantage in using phased antennas is that you can reduce noise and QRM pickup from unwanted directions to such a degree that you can hear everything on the band."

VE2HQ compared his home-brewed three-element Yagi with a sloper at 155 feet and found the Yagi to be superior in terms of signal strength. He also estimated the front-to-back ratio to be 22 dB.

With over 300 countries worked on 80 meters, **W4DR**, in comparing his four-element phased vertical array with other antennas, says he believes that the really high horizontal beams still have the edge. **N4AR** uses a pair of phased Bobtail curtains on 80

meters and finds that the F/B ratio on low-angle signals is often in the 15-20 dB range and the front-to-side is "astronomical." He prefers this antenna over Beverages for receiving as well. He also says that having an antenna with noticeable gain and a very definite controllable pattern has provided him with insights on propagation on 80 meters. He's probably one of the few to have worked Laos (XW8BP) on 80 from the East Coast . . . .

Attributing his success to a two-element delta loop array, **DJ0IA** says he's experimented with other antennas but keeps on coming back to it.

W1FV, showing true ham ingenuity, fit a well-designed and constructed phased vertical array into very narrow confines without any compromise in performance. He rates its overall performance on 80/75 meters as excellent. Winner of the single-band category in the 1984 *CQ* WW CW contest, he observed: "The vertical array is the best antenna I have ever used for the difficult paths (central Asia, Japan, deep Pacific, and Indian Ocean)." He has tried many slopers, high dipoles, and delta loops but finds the present configuration to be the best.

EA8ADP proves that sometimes a simple antenna can be a top performer if other conditions are met. His inverted vee has accounted for 225 countries worked on 80 meters in less than three years. He can normally be heard in the States when the rest of Europe is barely being copied. His inverted vee is supported at the 33-foot level on a tower atop a 148-foot high building situated on top of a hill.

propagation

As far as I'm concerned, this has to be one of the most interesting areas considered in the survey. The perspective gained by being able to review responses from different countries, continents — and in general, both ends of a communications path — starts to explain observations made over the years but not understood. I feel fortunate to be the compiler of this information and I'll try to convey some propagational correlations and leave the rest to you. A special vote of thanks is due to **SM6EHY**, who provided an extremely detailed account of his observations made during 20,000 hours of listening on the low bands.

160-meter propagation

For those not familiar with 80 and 160 meters, there is a tendency to lump them both together as "those noisy, short-range bands." That static exists on the lower frequencies is undeniable; that the communications range is limited is true sometimes, but not always. How else could one explain stations working over 300 countries on 80 and 200 countries on 160 meters, given the right conditions and activity? Furthermore, 160-meter enthusiasts are quick to point out that there's sometimes a world of difference between propagation modes on 160 meters and on 80 meters, and that one band should not be judged by the performance of the other. **K1MEM** best expressed this when he said "The basic creed is, don't go by 80-meter conditions to judge 160. 160 can be great to Europe or Africa when 80 is poor and vice-versa. QRN on one band is no indicator of the other. Also the beacons from Europe seem to have no relation to Amateur signals."

K5UR finds propagation to be best towards Africa and the Indian Ocean in the early evenings. At this time of day he's occasionally observed long path signals from Asia (YB, 9M, . . .) coming from the direction of South America. The same stations are also heard on a skewed path at a bearing of 235 degrees in the morning. Signals from Russia appear on a bent path with a heading of 75 degrees.

From the northwest, **KG7D** believes that chordal propagation is the mode on 160 when Europe is being heard at his QTH, yet not being copied at other locations in the United States. He believes this occurs when the F1 and F2 layers fail to merge. "It would be nice to know the optimum takeoff angle for chordal propagation. For me it appears medium angle north/south, low to the east, and again medium angle to the west. Frequent openings do definitely occur along the terminator lines, with signals coming from over the poles. Frequently we see enhanced propagation effects from the auroral zone."

VE7BS found a good opening occurring to VK almost every day throughout the summer of 1985. It started well before his sunrise and ended well afterwards. The observation of this path, he believes, is largely a function of the operating habits of Australians, pointing out that in June and July, VE7 sunrise occurs at a reasonable time in the VK evening. At other times of the year — for example, when VE7's sunrise occurs at 1430Z — the VKs have already gone to bed . . . thus no communications. But the path may still exist. This is somewhat analogous to a tree's falling in the forest; is a sound made even though no one's there to hear it? The answer is an emphatic *yes*.

SM6EHY echoed this same point in discussing the polar path between Sweden and stations on the East Coast of the United States and in the Midwest. Stations can be worked from 0800 to 1200Z on 80. Bjorn believes that the same conditions should hold true for 160 and attributes the lack of contacts to the low level of activity in the northern United States at this time. Bjorn also wishes United States Amateurs would improve their receiving capability; for example, on the 10th of June he failed to attract the attention of W9s after 15 attempts. He also feels that 160 can open up on occasions to a degree that has never been observed on 80 meters. For example, ZL3GQ was 599

+10 dB at 1652Z on December 24 for about 1 minute. The best he's been able to observe on 80 was 589 over both the short and long path.

80-meter propagation

Consisting of two DX segments separated by approximately 300 kHz, 80 meters sometimes also acts as if it were two different bands. Conditions can be excellent on the short path to Europe on SSB and poor on the CW portion and vice-versa.

More responses were received for this band than for 160 meters. Consequently the subject is divided into several sections: short path, bent path, long path and other influences.

• **Short path.** For the most part, those surveyed concentrated on the unusual paths either long path, bent, or skewed. Perhaps it's because the normal great circle path is in reality not all that common or at least certainly not as exciting. If one assumes that several "short paths" exist between two locations at the same time and one path is more severely attenuated than the other, then it's reasonable that the latter path (non great-circle) will be observed as the only path and classified as a bent path. For example, suppose a station in New Hampshire is in contact with a station in the U.K., and that they're both using single verticals. If at the time of the QSO the geomagnetic field is disturbed and the normal great circle path, at a bearing of 54 degrees, is severely attenuated, then signals coming from the second path, off the coast of North Africa, will provide the only viable means of communications. This situation, which has often been observed over the past several years from K2RR to the U.K., normally coincides with a fairly high K index. However, even the latter path disappears at still higher values of K approximately 24 to 36 hours after a major solar event. It's at these times that only signals from the South can be copied and usually only unilaterally. A case in point was during the recent Revilla Gigedo dxpedition, **XF4MDX**. It was as if a traveling blanket of absorption was moving from north to south. A few minutes after the fortunate contact, the signal from XF4 simply vanished. One could observe this effect by listening to which stations were able to get through W1 to W2 to W3, etc.

From his location in Louisiana, Mike, **NW5K**, is able to work into Africa between 0330 and 0630 and into Europe between 0600 and 0700Z via the short path.

Roy, **ZL4BO**, finds that the normal short path into Europe can vary as much as 30 degrees during his (southern) winter.

bent path

The term "bent path" can probably best be explained by saying what it's not: it is *not* a short or long great circle path. Can we then consider the vast number of other possibilities as bent paths? Were it only that simple! Perhaps it's more accurate to say that any path that's strictly not great circle, short or long, contains some element of bending. Remember, the distances covered in some cases are to locations more than halfway around the globe, where the time-varying ionospheric layers are in a different state of flux. Perhaps the signal starts out on a great circle path, only to be diverted by conditions along the way. Maybe this also explains why some of the paths appear to be unilateral or one-way; wouldn't conditions between the stations have to be the mirror-image of each other to be reciprocal? One way to analyze this effect scientifically might be through the use of simultaneous encoded transmission beacons, each with its own directional antenna at both ends of each path, with each transmitter keyed to an atomic clock. Precise time measurements would then determine the path interval and at least its length. But this, some might say, would take the fun out of Amateur Radio. Isn't it the unpredictability of the arrival of signals from exotic locations that makes DXing so exciting?

But I'm straying from the subject. Let's get back to which bent paths, under what conditions, have actually been observed.

Using a three-element Yagi, **W6RJ** observes that most northern European openings are bent somewhat to the south. Furthermore, he states that two stations with coinciding sunrise and sunset times can communicate 90 percent of the time, in the absence of high QRN levels.

ZL4BO attributes his success in bent path propagation to his extreme southern latitude, utilizing the path over both the North and South Poles when the normal (short path) and long path conditions aren't good.

SM4CAN notices that the best direction for working United States stations "short path" is at a bearing of 270 degrees as opposed to the indicated 330-degree great circle azimuth. During auroral conditions, he has worked W6NLZ with signals appearing to come from South America; this correlates with Peter's (W6NLZ) observation from his end of the path.

N4AR has observed signals from Japan, Hong Kong, and Indonesia from the Southwest. In fact, he always hears VS6DO and YB5ASO more strongly on this bent path than from the great circle direction. "This bent path is clearly in the direction of the terminator line [Gray-line — Ed.] and somehow these far eastern Asiatic and Pacific signals scatter into the terminator line path," he says.

Using his full-size three-element Yagi, **OH1RY** observes signals from Hawaii at a bearing of 90 degrees (east), whereas great circle calculations would indicate a northerly direct path.

W1FV, as well as many others on the east coast of

the United States, has observed the same bent path towards Japan and Southeast Asia in the morning by beaming to the southwest. Not only have other stations observed this path, but they've also seen it change abruptly (within a matter of minutes) from the southwest to the northwest when hearing Japanese stations. Using a rapidly switched phased vertical array, John finds that it often exhibits ill-defined directional characteristics for *long-haul* paths *around sunset* and sunrise, although the antenna is otherwise quite directional. "Perhaps this is explained by multipath propagation or some unusual radiation-angle and/or polarization effects?" he wonders, adding that "this ill-defined directionality is also noted on most signals during periods of a disturbed geomagnetic field."

long path

It's not uncommon to hear stations calling CQ long path around sunset from the east coast. What are they hoping to accomplish? And what's this phenomenon called "long path?" It's reasonable to assume that the practice first gathered momentum with stations on the higher HF bands — for example, on 20, 15, and 10 meters years ago. They observed that it was possible to contact stations by pointing their highly directional antennas in the opposite direction of the great circle short path. At the time very few stations on 80 meters had directional antennas and it was even less likely that *both* stations at opposite ends of the path would have directional arrays. That long path contacts occurred on 80 meters in those years was, however, certain; they just weren't called "long path." (I would like to hear from anyone who operated 80-meter long path years ago.)

N4RJ, W4DR, K2FV and **W1FV** have been able to work into western Australia, Indonesia, and Japan at their sunset by beaming southeast, utilizing primarily the "long path."

From the other side of the path, **VK6LK** finds that his long path opening to the East Coast of the United States normally starts about 15 to 20 minutes after his sunrise and lasts 20 to 30 minutes when the conditions are good. "My first long path opening was with W1FC on 13 September this year, and I would expect the path to close sometime next May. One problem here is short skip from YB. They can drown the weaker long path signals from the United States." From Japan, **JA1FRE** has also noticed the existence of a US-JA long path through his (reception) observations of east coast United States stations at 2130Z.

Other, long path shots not well known by US stations of course still occur. **JA1FRE** works into the UK and Scandinavian countries at 0800Z, taking advantage of this mode. **SP3GEM** and other European stations are able to work into the Pacific areas using the long path. **K5UR** has a morning long path into Europe that isn't too common, but when it does occur, produces high signal levels.

Using his phased Bobtail curtain, **N4AR** has shown how reliable the early morning long path into central Asia can be. He has maintained a schedule over the years with **UL7GW**, talking to him approximately *100 times per season* using this mode. "During that period, many other Asiatic Russians were easily worked on a daily basis," he wrote. Besides these USSR stations, he also worked **XW8BP** and other southeast Asians at this time of day.

What's even more interesting is that he found that the path essentially disappeared during the sunspot *minimum* of 1974-1975. This is contrary to the widely accepted belief that sunspot *minimum* years provide the best low-frequency DX. N4AR commented, "While stronger signals (during this period) were evident from Europe in the evening and an occasional UA9 or UL7 QSO was possible in the evening, for all practical purposes the morning long path disappeared."

As stated earlier, this path returned during the most recent solar maximum. Fortunately at that time, **HS1ABD** and **UL7GW** were again active, serving as markers for band openings.

N4AR remarked that with the phased Bobtails, "There's no question deciding if signals are indeed coming from the direction of the long path. This tends to be along the terminator line during the 1130 to 1230Z window. Looking at the MUF limited F2 hops doesn't seem to explain the loss of signal (during sunspot minimum) adequately. In any event, it's certainly a real phenomenon, at least from this part of the world."

SM6EHY disagrees with the contention that a true long path condition exists with signals *originating* and *terminating* in the same hemisphere. "Long path — with the meaning of true long path direction — does not occur on 80 meters," he wrote. Signals arriving at SM6EHY's QTH from KL7, KH6, VE7, W7 and W6 have the same azimuth angle: 110 degrees. "They arrive at a very high angle (50 to 60 degrees), with some noticeable echo. This holds true only for the same hemisphere. Signals coming from the southern hemisphere have all true azimuth angles for both short and long paths." He also believes that at W0, W6, and W7 sunrise, stations should not be beaming long path but using the Arctic path instead.

other influences

Through his 20,000 hours of low-band observations, **SM6EHY** credits the significant influence of the aurora on propagation. "The conditions this far north are very heavily affected by aurora with regards to paths to stations in the northern hemisphere such as JAs and Ws. The maximum particle radiation from the sun (which causes the aurora) is predicted to occur ap-

proximately two years after the sunspot maximum. In this last cycle the aurora has consistently occurred from December, 1981, until now (March, 1986) leaving the polar path open only on special occasions." When this polar path is open SM6EHY can work the east coast and midwestern United States from 0800 till 1200Z. Japanese stations can be worked 24 hours a day. This occurs from mid-November until mid-February, when the path is open.

where is this auroral activity?

SM6EHY continued, "The aurora expands between Spitzbergen (JX) and northern Norway (LA) and under most conditions the maximum activity is experienced close to 64 degrees N (latitude). From 144 MHz observations, the reflection belt travels from the far east through the north to the west and then disappears. Occasionally, it's the other way around.

Bjorn went on to mention that the aurora can change very rapidly — within a period of seconds — virtually closing down a good path. When the auroral activity is *south* of his location the "arctic is like an open field where signals experience very little attenuation."

what other paths does it affect?

"A bent path has been observed during auroral activity," Bjorn wrote. "The signals coming from W1 [stations] just cannot travel through the arctic region [the auroral can be considered to be a 0 to 30 dB pad, depending on the level of activity and signal path — Ed.] and must be reflected in an area close to the equator and other places at the same time. This causes the signal to be heard with some echo and QSB."

benefits of a low-angle radiator

With his phased vertical array and large ground system, Bjorn feels he's able to launch an extremely low-angle signal that can go *under* the aurora's attenuation belt. "The lower you go in frequency, the more important is the takeoff angle; the angle fluctuations are more pronounced at lower, rather than higher, frequencies," he said.

Bjorn believes it's important to transmit at angles between 0 degrees and 10 degrees on 80 meters. "A low takeoff angle antenna can be used very effectively during auroral activity to contact some stations at least *part of the way* through the auroral region. A case in point is when he's not able to hear the W1s, but can still work OX or VE8 stations with good signal strengths."

ducting

Another propagation mode noticed by **SM6EHY** and others involves signals entering at one location, traveling and being trapped between ionospheric layers, and finally exiting at a second location, whereby stations in-between cannot copy either end point. Specifically, on north-south paths signals from stations as far south as 15 degrees follow normal paths. Further south, ducting appears to occur. There are times when Swedish stations north of SM6EHY are able to communicate with South Africa but not be heard by him.

Ducting also possibly occurs on the North Atlantic path between the east coast of the United States and Europe. This might explain the high signal strengths received across the Atlantic by stations using low dipoles that launch high-angle signals. If it were simply a case of multihop propagation between the ionosphere and earth, signal levels would be lower. However, reflections *between* layers, on the other hand, would account for less loss. The entrance and exit requirements for ducting might be high-angle.

when are low-band conditions best?

SM6EHY echoes the sentiment expressed by N4AR that low-band conditions are probably best during the period of maximum solar activity (sunspot maximum). He explains that it's probably not noticed, since proportionately fewer stations operate on the low bands when the MUF is high and 10 through 20 meters are more heavily utilized at that time. It's possible to test this hypothesis during the next sunspot maximum, because there'll be more stations on than ever before using high-performance directional 80-meter antennas.

disturbed geomagnetic field affects path

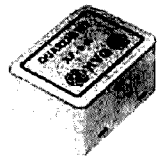
N4AR kept close track of WWV's A and K indices for many years and believes that neither the bent path nor the great circle long path appear to be beneficially influenced by geomagnetic disturbances. However, with the onset of a geomagnetic field disturbance, spectacular increases in signal strength are occasionally evident over the short path. Also observed by others, the latter situation is normally a precursor of the arrival of the lower energy electrons from the sun 24 to 36 hours later and its associated high level of D layer absorption.

plans for the future

Because improvements are always possible in any station — big guns not excluded — the final question asked in the survey was "What improvements are you considering making to your station?" The responses to this question included ideas for improving receiving capability, installing higher performance transmitting arrays, increasing tower height, using a bigger radial system, installing lower loss transmission line and even acquiring a better site.

• **Receiving capability improvement. K3ZO** wants to improve his 160-meter reception in general. **N1ACH**

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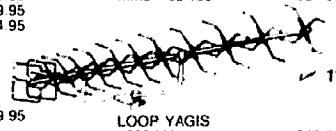
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wants to lengthen his 160-meter Beverages from 1.5 to 2.0 wavelengths. **4X4NJ** looks forward to an improved R4C for 160 meters. **SM4CAN** wants to use better Beverages. **N4RJ** feels that his Beverages could be better tuned and that adding a preamp at the antenna would help. **K1MEM** will terminate his Beverages and run additional ones SE and SW.

• **Transmitting array improvement. VE3BMV** wants to put a two- or three-element quad on his tower. **OH1RY** wants to install a two-element shortened 160-meter rotary. **OE6MBG** would like a two-element driven IV beam; **K2FV**, an additional half square. Phased verticals or phased delta loops would please **SP3GEM. KG7D** plans a three-element 160-meter parasitic vertical array using shortened elements. **JF1IST** would prefer a four- or five-element 80-meter rotatable Yagi. **DL0WU** wants a three-wavelength on a leg 80-meter rhombic; **K5UR**, a two-element phased 160-meter vertical aray. **YU7PFR** would like to construct four- or five-element wire Yagi, and **SM6EHY** a four-element 160-meter phased vertical array. A three-element full-size 80-meter Yagi at 160 feet would interest **VE2HQ; JA1FRE** would enjoy a two-element full-size Yagi. **W4DR** wants an 80-meter horizontal beam at 150 feet. **W6RJ** plans to raise his existing 80-meter three-element Yagi on his tower.

Other items on the big-gun wish list include increasing tower height (**PA3DFU**), adding more radials (**W3BGN** and **W1FV**), and using lower loss cable; **W1NH** wants to switch to hardline coax to reduce losses on his 500-foot run to the antenna. **WB2ITR** expresses a commonly felt need for a new site, one that has more land and better topography.

conclusion

Well, now we have it: the secrets of successful low-band operation. Well, not quite. One necessary ingredient that's hard to quantify is, of course, the desire to succeed — in this case, to have the very best signal on the band. Besides the other three factors mentioned before (hard work, time, and expense) there's also the dream, the enthusiasm that keeps you climbing to greater heights, laying out just ten more radials and extending that Beverage even further, hoping that one night as you're scanning, that BY4 will come pounding through on 3.505 with a never-before-heard clarity and strength.

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ham radio TECHNIQUES

Bill Orr W6SAI

a wrangle over Wrangell

Want to work a "new country?" Maybe you can. Unknown to most radio Amateurs, the United States has claims to Wrangell Island (UA0) and other island territories north of the Soviet Union. The situation isn't clear, and maritime-boundary discussions about this region — spurred on by the discovery of oil — are underway .

In a recent editorial *The Wall Street Journal* called attention to deliberate Soviet harassment of U.S. drilling rigs in the Navarin Basin. This harassment, which includes dangerous overflights by Soviet aircraft, is one of several subjects of discussion between the two powers. Although the issue bears watching, only limited information about it has appeared in the daily press.

Little-known Wrangell Island seems to be assuming an importance greater than its size. I doubt if there will be much radio Amateur activity from that part of the world until the boundary issue is settled.

the five-band coffee pot

Mobile antennas for the HF bands have always been a problem because of low operating efficiency and restricted bandwidth. A mobile enthusiast, Varouj Kalinian, (N6DBH, ex-OD5CS and F0HN) has solved this problem to his satisfaction. His solution is to use a short mobile antenna, but to increase the diameter of the element below the loading coil (**fig. 1**). For the base section, he used an old coffee pot about 8 inches (20 cm) in diameter and 14 inches (35 cm) tall. His loaded whip sections were mounted to the top of the pot (**fig. 2**).

Varouj removed the electrical components of the pot, drilled a hole in the middle of the plastic cover, screwed the cover to the pot, screwed a 40-meter Hustler resonator and whip to the top and then adjusted the antenna to resonance. He found that

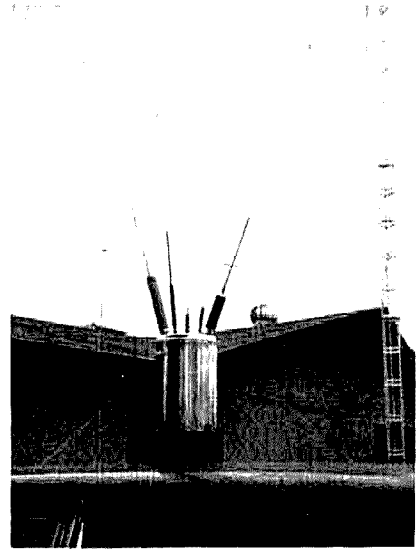


fig. 1. The "coffee-pot" mobile antenna of N6DBH mounted atop his station wagon. Separate resonators are used for various hf bands.

he had a bandwidth of 150 kHz between the 1.75-to-1 SWR points on the feedline on 40 meters.

Encouraged by the results, Varouj added other resonators and antenna tops for the other HF bands. The results were a bandwidth of 60 kHz on 80 meters, 200 kHz on 20 meters, 300 kHz on 15 meters, and 800 kHz on 10 meters. In a short time he had a "five-band coffee pot antenna."

Varouj admits the antenna looks a trifle bizzare and that he gets curious stares from other drivers. He reports that one driver pulled alongside, handed Varouj a coffee cup, and yelled, "Fill me up!"

a mini-antenna for 160 meters

It's unfortunate that a halfwave antenna for 160 meters has to be so big. Walt Bollinger, AF3V, must have thought about this as he tried to put out a reasonable signal on that band from a small lot. His solution to this problem is shown in **fig. 3**. This configuration is a loaded Marconi antenna whose overall length is about 50 feet (15 meters). It is mounted about 20 feet (6 meters) above the ground. A 6-foot (1.8 meters) ground rod is used in conjunction with a 65 foot (19 meters) long counterpoise wire running beneath the antenna, just below the surface of the soil. As illustrated, the antenna operates over the range of 1.8 to 1.825 kHz with an SWR figure of 1.5-to-1 or less. If a variable capacitor is connected in series at the feedpoint (X), the antenna will operate over the rest of the band with low SWR. The approximate capacity necessary to achieve this without exceeding the 1.5-to-1 SWR is as follows: 1.85 to 1.92 MHz, 700 pF; 1.92 to 1.97 MHz, 450 pF; 1.97 to 2.00 MHz, 350 pF.

Walt uses a three-gang variable capacitor long ago removed from an old tuned-RF broadcast receiver (QRP only). It has three 365 pF capacitors on a common shaft. "For its size," Walt

says, "I think this antenna will surprise most everyone. It seems reasonably omnidirectional and provides me with a lot of fun on the 'top band'."

a rectangular loop antenna for 80-10 meters

Walt, AF3V, checks in with a second interesting multiband antenna **fig. 4**. A rectangular loop in the vertical plane, its overall length is about 58 feet (17.6 meters) with the bottom wire about 7 feet (2.13 meters) above ground. Its unique feature is the knife switch (SW) placed across the insulator in the center of the bottom wire. Easily reached, the switch permits the operator to change from an open loop to a closed one quickly. The feedline is a home-made, open wire line about 42 feet (12.8 meters) long. It drops down vertically to within 8 feet (2.4 meters) of the ground and then extends horizontally to the station, perpendicular to the plane of the antenna. The antenna is adjusted to resonance by means of an old Johnson "Matchbox" tuner. The antenna switch is closed for operation on 7, 10, and 21 MHz and is open on the other bands. With a different length feedline, different switch settings may be necessary.

Walt supports the center of the top section to offset the weight of the feedline. He says that the antenna outperforms the center-fed horizontal wire that he had previously and reports that the pattern appears reasonably omnidirectional on all bands.

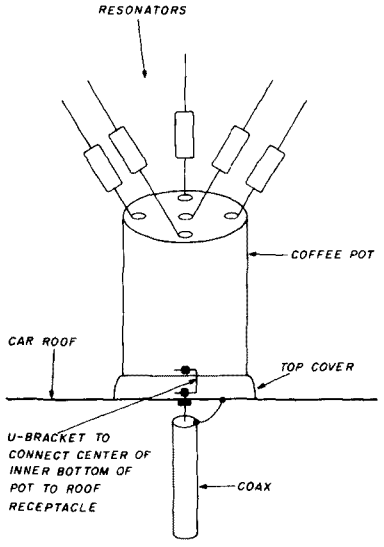


fig. 2. Coffee-pot is screwed to cover which acts as a base insulator. U-bracket fixes bottom inner section of pot to existing VHF antenna base in middle of station wagon roof.

a simple antenna for 40 and 30 meters

Many Amateurs say they enjoy the 30-meter (10 MHz) band because there are no disconcerting contests on it and because it's possible to rag-chew with a DX station without having a horde of eager beavers making rude noises on the frequency, unhappy that they're thwarted in their search for a "new one."

Harvey Hunter, W8TYX, enjoys 40- and 30-meter operation with the simple antenna shown in **fig. 5**. Basically, it consists of a 30-meter ground plane antenna, roof-mounted, with a ground screen of 60 radials. The radials also serve with the 40-meter antenna, which is mounted a few feet away from the 30 meter whip. About 5 feet (1.5 meters) of the 40-meter antenna is vertical; the remainder is more or less horizontal, about 25 feet (7.62 meters) above the ground.

Slightly longer than a quarter-wavelength, the 40-meter wire antenna has a feedpoint impedance of close to 50 + j50 ohms. A good match to the 50-ohm transmission line is achieved by using a series capacitive reactance of −j50 (450 pF) between the feedline and the base of the antenna.

The 30-meter antenna has a feedpoint resistance of about 35 ohms, so a simple L-network is used to match it to the transmission line. The antenna adjustments are slightly interactive, but tune-up using an SWR meter quickly compensates for that.

the "hy-tower" antenna operates 160-10 meters

The many Amateurs who have the "Hy-Tower" vertical antenna may find the automatic tuning arangement for multiband operation used by Paul Scholz, W6PYK, of interest. The scheme should work well for any vertical antenna over 35 feet (10.7 meters) high.

The basic network, shown in **fig. 6**, consists of a series-connected LC circuit (L1, C1) in parallel with a second inductor (L2). The series network is adjusted for minimum SWR on 75 meters with the parallel inductor set to maximum value. The last step is to adjust the parallel coil for resonance on the 160-meter band. The two adjustments don't interact if the 75-meter adjustment is done first. Operation of the antenna is normal on the higher frequency bands.

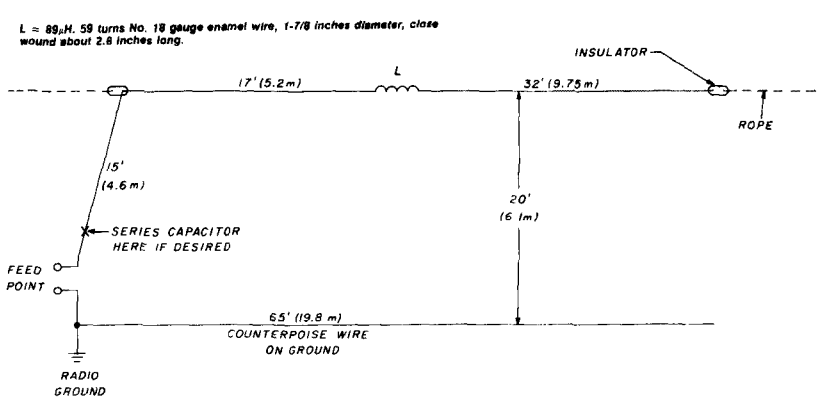


fig. 3. The compact 160 meter antenna at AF3V.

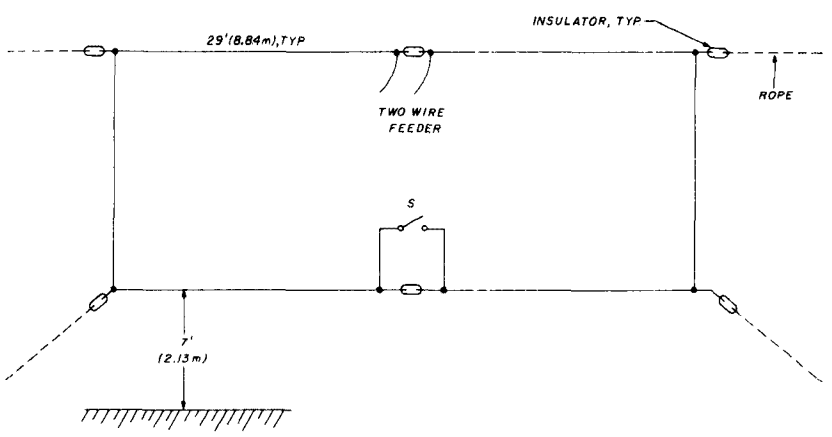


fig. 4. The multiband loop at AF3V. Switch permits operator to change from open loop to closed loop. Loop operates on all bands between 3.5 and 29.7 MHz.

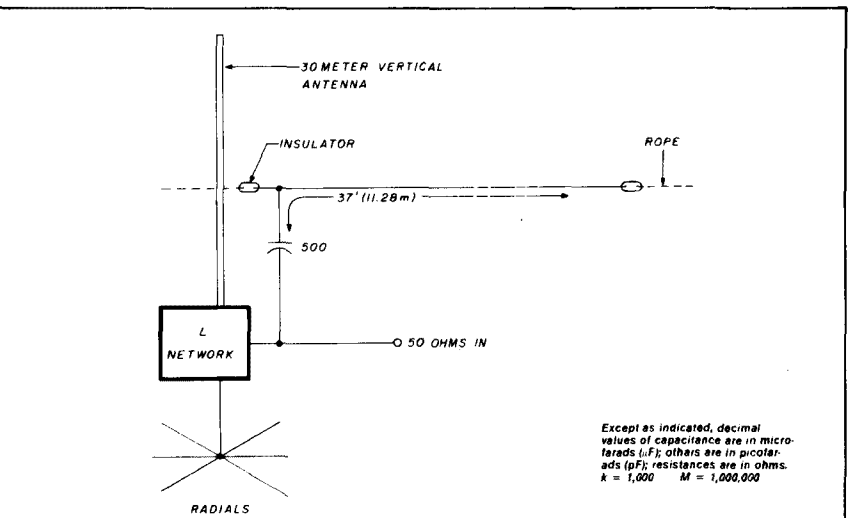


fig. 5. Two-band antenna for 40 and 30 meters used by W8TYX.

For 75-meter operation, a range of 200 kHz is obtained between the 2-to-1 SWR points. On 160 meters, operational bandwidth is about 50 kHz.

Although Paul recommends the component values shown in **fig. 6**, he has a computer program that provides component values for different LC ratios in the network. For low-power operation, 5 kV transmitting-type mica capacitors can be used; for high power, he recommends a variable air capacitor with 0.125 inch (3.2 mm) spacing.

a rugged, long-life quad

Many exasperated Amateurs have wondered how it's possible to keep the Quad antenna alive and well year after year. The common weak point of many Quads is the junction between the wire and the Quad arm; the wire tends to either break at this point or "saw" its way through the Quad arm.

In my February, 1986, column I suggested one way of easing this vexing problem. I recently received a communication from Paul Atkins, K2OZ, suggesting a different (and simpler) technique that has permitted his Quad to stay aloft for over 13 years. On the advice of a friend in the wire business, Paul rebuilt his Quad with silverplated teflon-coated wire. The teflon coating sheds water easily and retards ice buildup. The manner in which he attached the wire to the Quad arm is shown in **fig. 7**. The teflon covering is removed from the Quad wire about 10 inches (25 cm) out from each arm. The slug of teflon removed is about 2 inches (5 cm) long. A thermal stripper is recommended. Next, two pieces of No. 14 tinned copper wire long enough to be twisted in opposite directions are tightly wrapped around the wire element. Using a large iron, the ends of the twisted wires are soldered to the stripped portions of the antenna element. Run solder over the connections to stiffen them. This brings the moment arm away from the spreader arm and minimizes flexing at the point the element is secured to the spreader. Finally, the flux is wiped off the connections, which are then wrapped with

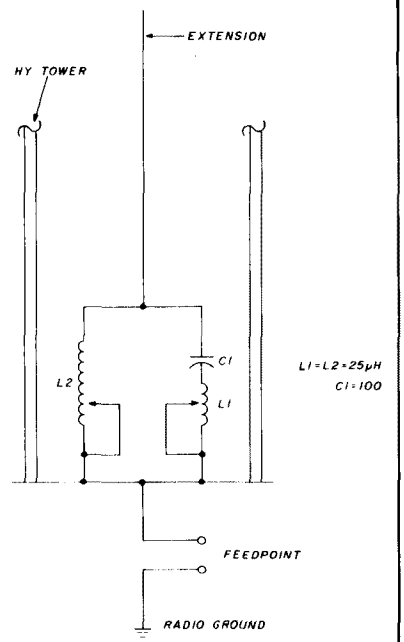


fig. 6. The two-band network used by W6PYK for automatic 160 meter operation of his "Hy-tower" antenna. Network operates as an electrical switch.

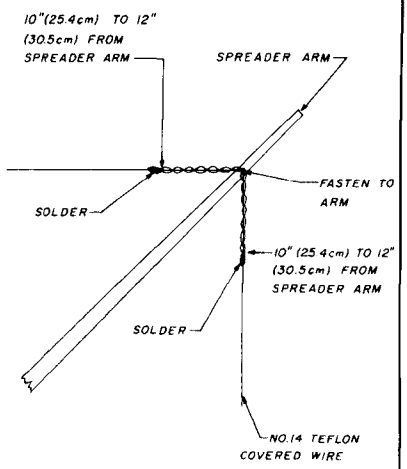


fig. 7. Quad wire is strengthened by twisted No. 14 wires wrapped around it and soldered about 10 inches to 12 inches from spreader arm. Installation at K2OZ has been up for over 13 years.

tape to protect the exposed soldered joint from the weather.

The final addition to Paul's Quad was to tie the driven element to the reflector element with polyethylene cord. The tie is made near the tips of the spreaders to keep the arms from distorting too badly in high winds or under heavy ice loading. The black polyethylene cord — about 1/8 inch (0.32 cm) in diameter — is still in service.

the 40-meter DX beam at K6NQ/7

Forty-meter DX operators have heard the strong signal of Jim Villasana, K6NQ/7. This is a tough band for serious DX competition because some of the "big guns" use multielement Yagi and Quad beams. Great antennas, if you have the time, money and space to put 'em up!

Jim's solution to the 40-meter DX problem is to use two sloper dipoles hung from a 140-foot (42 meters) fir tree and fed 270 degrees out-of-phase (**fig. 8**). The slopers are at a 45 degree angle with respect to ground. The centers of the dipoles are at the same height. This places the top ends of the dipoles at 100 feet (30.48 meters).

The dipoles are fed with equal lengths of RG-8X coax (230 feet/70.1 meters) and the phasing of the reflector element is accomplished by adding an extra 3/4 wavelength of line (approximately 86 feet, or 26 meters) coiled up inside the operating room.

While Jim has tried various other antenna arrangements, he finds that this simple antenna works best for DX. Dismayed to find that the array exhibited little gain on short-haul contacts, he was amazed at how well it worked on long-haul DX. He found, to his gratification, that his 7-MHz signal was as strong in Europe as those of other west coast DXers running high two- and three-element Yagi beams. On very good openings, the simple

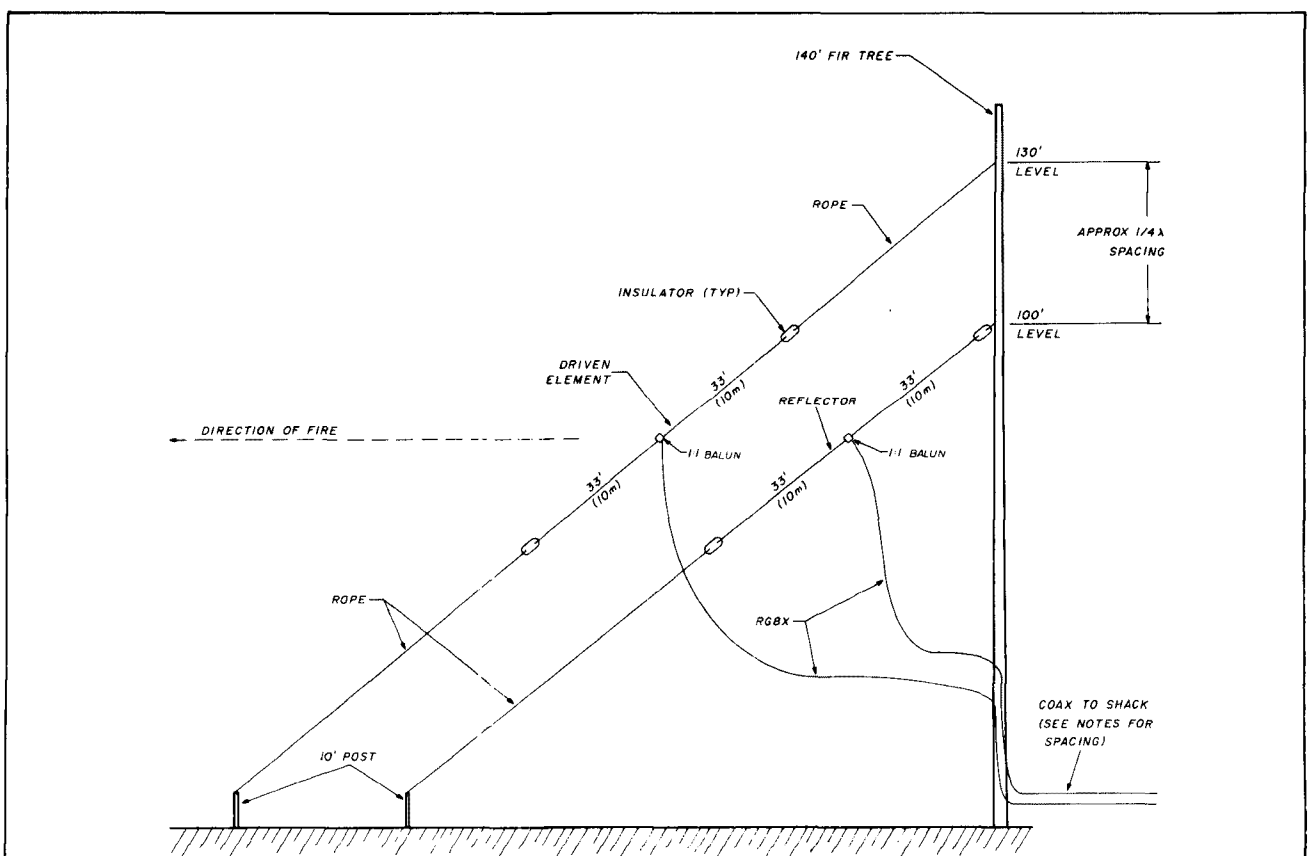


fig. 8. 140 foot high fir tree in yard of K6NQ/7 provides support for "hot" 40 meter beam!

sloper antenna was better than 1 S-unit over competitive Yagi antennas located at the 120-foot (36.6-meter) level.

Its beam pattern is broad enough to cover Europe and Northern Africa; Jim estimates about 90 degrees beamwidth to the −3 dB points. He tried several phasing schemes to get the required 90-degree phase difference and found out the hard way that the best results were obtained when the rear element used 3/4 wavelength phasing.

Jim wishes he had the room to string up two or three arrays of this type to get full coverage, but that will have to wait until he has more real estate. Meanwhile, he does a fine job with a very inexpensive antenna!

the British "zip code"

Some time ago I admitted that the British Postal Code was a mystery to me — and to most other Amateurs as well, I suspect. Jim Miller, G3RUH, very kindly sent me a Post Office brochure explaining the system in such a lucid manner that even I could understand it. (The Canadian code works on the same principle, I am told).

In brief, the code is a simplified address, with each part of it focusing on a progressively smaller area. The United Kingdom is divided into 120 areas, each being identified by the first two letters of the code. As an example, the code MK42 8LA is given. The MK indicates the Milton Keynes area of England; the 42 stands for a district within the Milton Keynes area; the digit 8 indicates a sector in district 42; and the LA designation identifies the street. The code is read by a machine which converts it to two rows of blue dots imprinted on the envelope. The dots, in turn, are read by a mechanized system which sorts it on its way. Thus, the code helps identify an individual person on a particular mail route! It's also often imprinted on bicycles, automobiles, and other vehicles to assist in the return of stolen property.

amateur service bulletins

The popular "moonbounce" notes have been reprinted and are again available. For a copy, send five first-

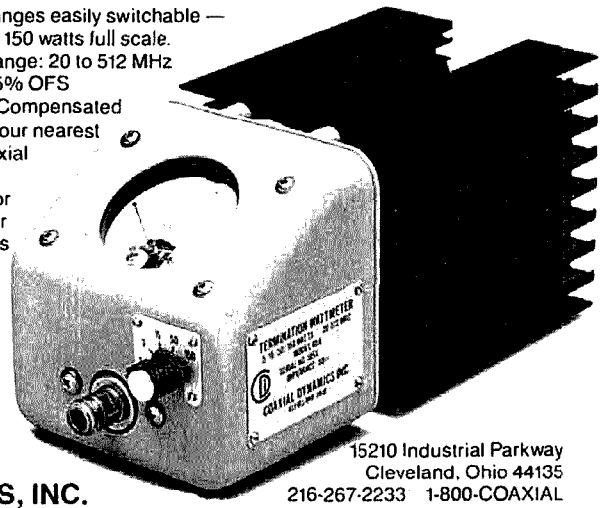
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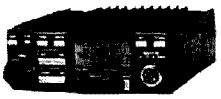
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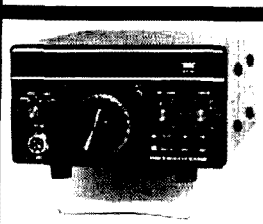
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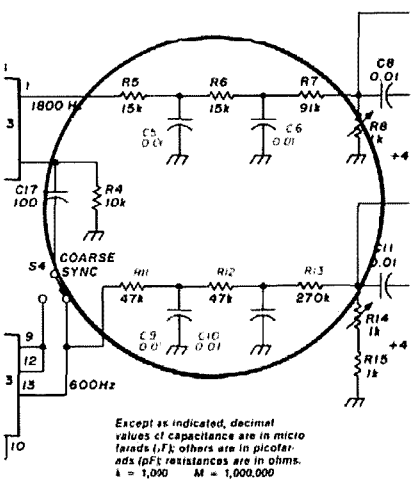
In WB4EHS's March article, "Build Narrowband RF Filters (page 10), the word "smaller" in the second sentence on page 14 should be changed to "larger."

In the program listing on page 15 (**fig. 7**), **line 145** should be corrected to read as follows:

```
145 INPUT O:IF O<1 OR O<5
    THEN GOTO 145
```

signal generator

Values for four capacitors and labels for three resistors were omitted from **fig. 8** of YB9ATA's article, "Two-tone Signal Generator" (February, 1986, page 26). The corrected portion of the schematic is shown below.



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

make a homebrew sheet metal brake
for chassis construction projects

Save time, money — and build a better box

Radio Amateurs have been bending aluminum at home for years, making brackets, panels, and boxes for electronic construction projects. Light weight, workability, good corrosion resistance, and electrical conductivity make aluminum an ideal material for small hand-fabrication projects. But classical bending techniques that require a hammer, vise, and pliers can result in dents and scratches, and extra work straightening undesired bends.

This article describes how to make a sheet metal bending tool that makes quality bends in thin metal without damage. The tool is made with just a few dollars worth of steel angle stock and some common household hardware.

Because the correct choice of material is important, information is included to enable you to identify aluminum alloys for bendability, strength, and corrosion resistance. A construction project of a small box is included as an example.

bending theory

Figure 1 shows the physical arrangement required to bend sheet metal without causing deformation in areas other than the bend. The bar holds the workpiece to the table while the leaf forms the corner around the bar's "knife edge." The major objective illustrated in **fig. 1** is to maintain a precise hinge centerline between the moveable leaf and fixed table at the bend corner. **Figure 1B** shows the direction of maximum bending force at the completion of the bend. The leaf must be rigid and the bar must be held securely to maintain a consistent bend radius along the work.

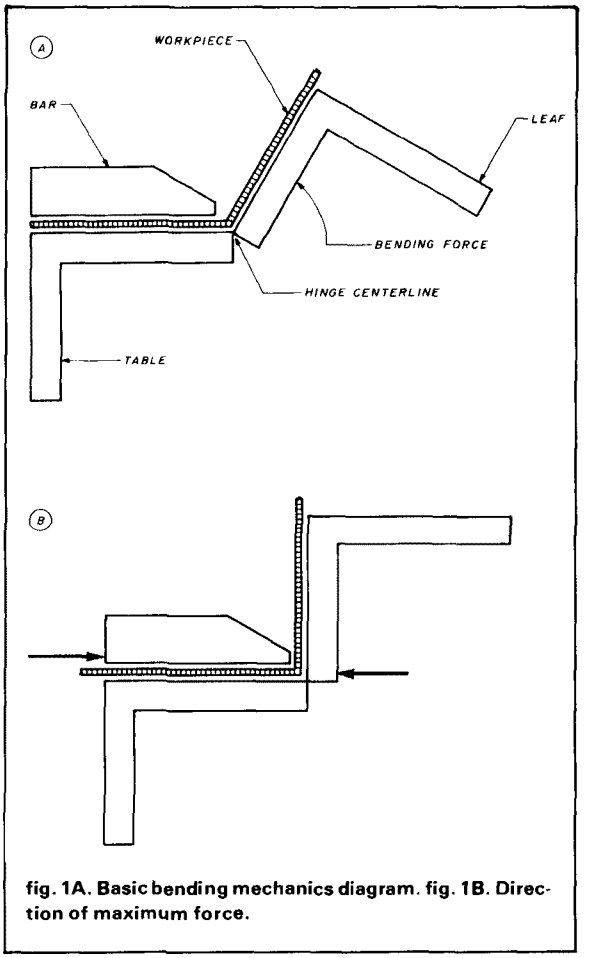


fig. 1A. Basic bending mechanics diagram. fig. 1B. Direction of maximum force.

By Cliff Klinert, WB6BIH, 1126 Division Street, National City, California 92050

table 1. List of aluminum alloys with zero bend radius (wrought sheet 0.062 inch (1.6 mm)-thick cold bent, 90 degrees)

Alloy	Yield Strength (psi)	Corrosion Resistance
1100-0	5000	A
1100-H12	15000	A
1100-H14	170000	A
2014-0	14000	D
2024-0	11000	D
3003-0	6000	A
3003-H12	18000	A
3003-H14	21000	A
5005-0	6000	A
5005-H12	19000	A
5005-H32	17000	A
5005-H14	22000	A
5005-H34	20000	A
3004-0	10000	A
3004-H32	25000	A
3004-H34	29000	A
5050-0	8000	A
5050-H32	21000	A
5052-0	13000	A
5652-0	13000	A
6061-0	8000	B

Note: Corrosion resistance is given in decreasing merit from A to D. Metal with an A or B rating can usually be used outdoors without protection. A D rating indicates that protection is required to prevent corrosion. Data was compiled from references 1 and 2.

We will later see that this can be a limiting factor in bending thicker or stronger material.

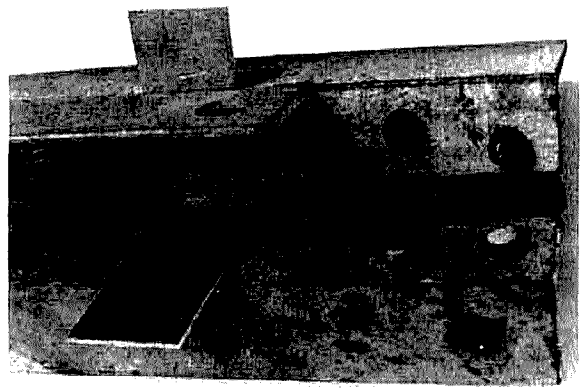
selection of alloys for bending

Availability of the correct material may limit the practicality of homebrew sheet metal projects. To determine which materials will be bendable, it helps to understanding the use of *alloy numbers*. The most common system of alloy numbers uses a four-digit alloy number followed by a combination of a letter and numbers that indicate the temper or hardness. The two major types of hardening are the "H" (strain hardened) and "T" (thermal hardened) processes. In the strain hardened system, the H is followed by two or more numbers, the second of which indicates the relative hardness of the material. The T in the temperature hardening system is followed by one or more numbers, the first of which indicates the relative hardness. If the four-digit alloy number is followed by an "0," the material has been annealed to make it as soft as possible and is readily bendable. The 0 material is soft for easy forming with the possibility of being hardened later. Incidentally, material marked "alclad" is covered with a thin layer of relatively pure aluminum for extra corrosion resistance.

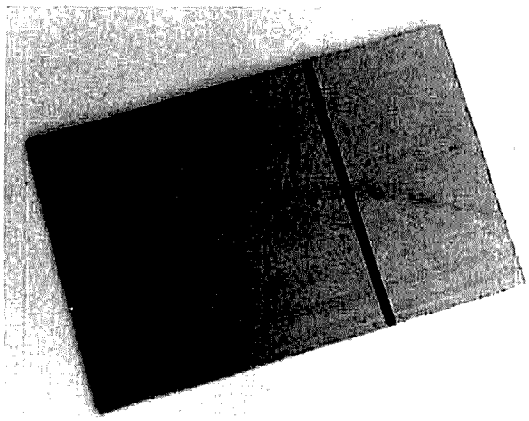
The most desirable alloy used for bending and forming is 3003-H14, which combines optimum strength and workability, making it the ideal material for box and chassis construction. The "3003" indicates that this series of alloys is strain hardened without any subsequent heat treatment. The first temper digit, "1," indicates strain hardening without any subsequent heat treatment. (Another digit in the first position, such as a "3," could be used to indicate a low-temperature thermal "soak" to improve ductility.) The "4" indicates the relative hardness. A higher number denotes a harder material.

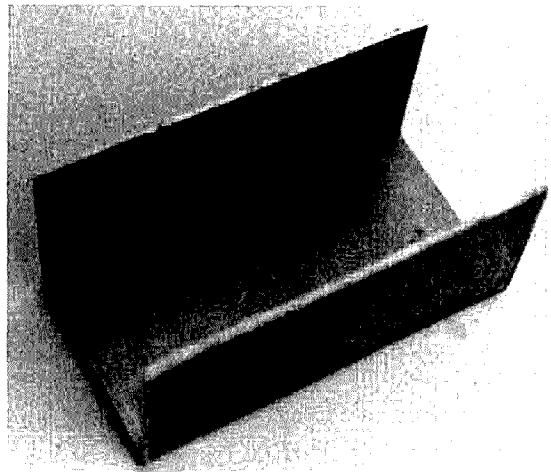
Table 1 is a shopping list of alloys appropriate for bending[1,2]. Since low cost is almost always a consideration, the best source of materials in small quantities will usually be a salvage or surplus store. Aluminum manufacturers typically require a $300 minimum purchase, and hardware store material is usually quite expensive in the small quantities required. For best results, take a pair of gloves, a tape measure, and a copy of **table 1** to all the metal salvage yards in your area. Prices should range from 75 cents to $1.50 per pound. Sheets are marked in alloy numbers (as shown in **table 1**) and thickness in thousandths of an inch. The optimum thickness is ".040" (1.02 mm). If a sheet is not marked, try to bend a corner with your fingers to determine bendability. You'll usually be expected to buy the entire piece, and cutting service may or may not be provided. Cutting may cost from 50 cents to a few dollars, so you may wish to take some sort of cutting tool with you if you want to buy a piece too large to carry home uncut.

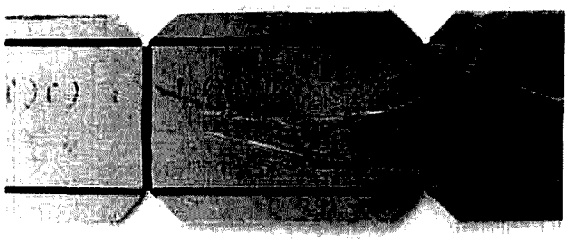
Salvage steel angle stock for making the bending tool should be available for ten to 25 cents per pound. The type of industries in your area will determine the type of salvage material available. Please note that "3003-H14-.040" is the optimum material for our projects. Thicker or harder material may produce bends with larger radii and make more crude-looking parts.

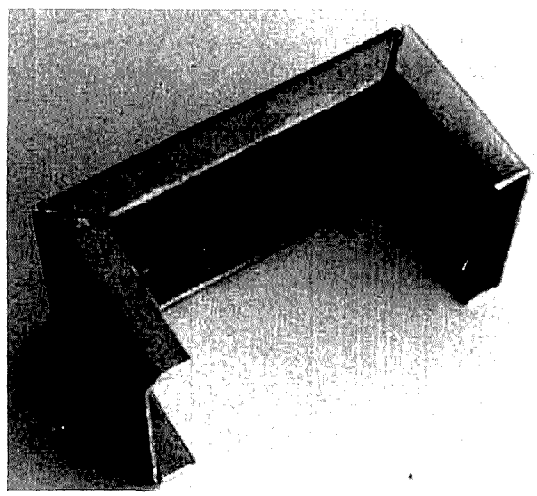


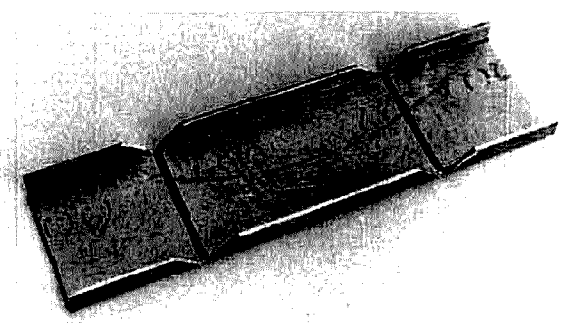
The construction of the bender, location of the hinges, attachment of the bar with bolts, and a small piece of sheet metal being bent.









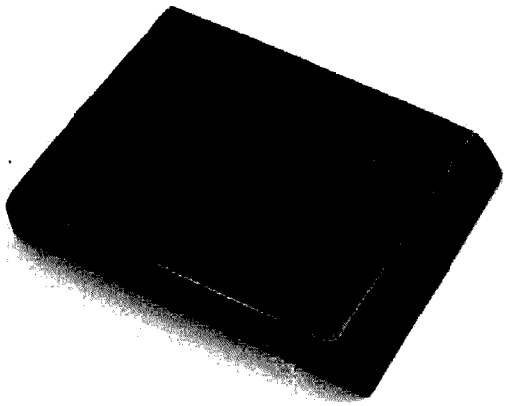


The bend sequence for making a small box. The material was 6061-00.062 thick, somewhat too thick for making precise bend corners. Note that a small hole was drilled at the intersection of the double bends to allow space for the material thickness.

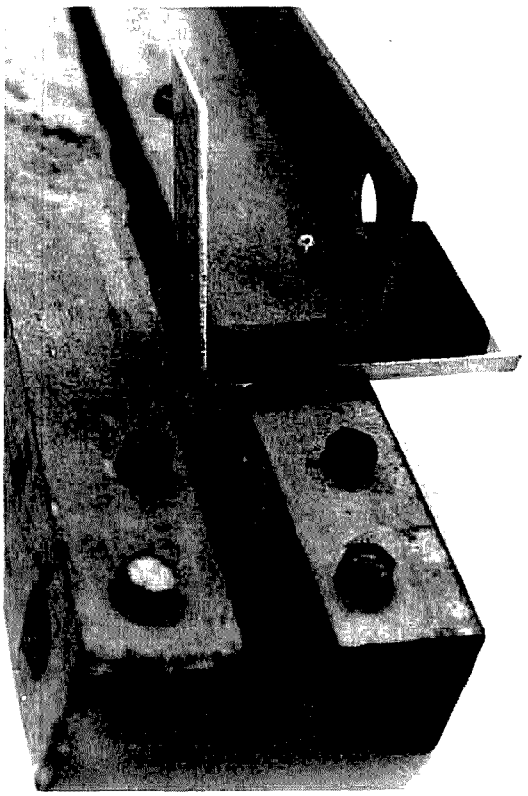
building the bender

The photographs show how the bender was assembled from angle steel and two door hinges. The overall width was about 3 feet (1 meter). You can select dimensions appropriate to the material available. The wider the bender, the more likely it is to warp in use. The photographs were taken to illustrate the geometry of the bender as shown in **fig. 1A**. Depending on the shape of the hinges used, the ends of the angle stock must be notched for clearance. In shimming the hinges with washers, locating the hinge centerline required some patience.

The corners were cut with a circular saw using a metal cutting blade (i.e., a carborundum disc). Grinding or filing the angle steel edges along the hinge centerline to produce a square corner is optional since even a fairly crude tool will produce good results with the right material. The "bar" is another piece of the angle steel bolted to the table. The bar could be secured with C-clamps, but the bending force shown



This piece was cut from a steel bar to allow making a double bend.



The die can be used to accomodate a part that has been bent in the opposite direction. The edge of the die is shown for clarity, but would normally be mounted between the two bolts holding the bar to the table.

in fig. **1B** may cause the bar to slide away from the leaf when you bend stronger material. You may wish to reinforce the leaf by bolting another piece of steel behind it, if you have problems with warping later.

Clamp the table in a vise, attach Vise-grip™ pliers to the leaf for handles, and the bender is ready for use.

making a box

Obtain a sheet of bendable alloy and cut some pieces for practice bends. Mark bend lines on the aluminum as shown in the photographs. Some practice will be necessary to determine where to place the material relative to the bend line and to anticipate the amount of springback in the bend angle. You can use tin snips or a nibbler; most wood sawing methods (using power tools) can be applied to aluminum as well. A bandsaw is ideal for thin material, and a circular saw with a plywood cutting blade is excellent for thicker sheet. A router works well for trimming or shaping, and most of the bendable alloys can also be whittled with a pocket knife. Use paraffin to lubricate the blade, wear ear plugs, and observe the normal safety precautions.

The photographs illustrate the sequence of bends for making a small box. A die was made from a piece of steel bar with an angle cut for a knife edge. (The size of the die determines the size of the box to be bent.) The die was used as a "finger" in a pan brake to make a bend perpendicular to a previous bend. The bar in a pan brake is made of several removeable fingers in various widths to allow for making double bends like this.

It is common to etch the finished part in a lye and water bath to produce an attractive finish. If lye is not available, welding supply stores sell aluminum cleaning solutions that work faster, but are somewhat more expensive. A quart bottle of cleaning solution costs about $5, but it can be diluted in water for several uses. Be sure to follow the safety instructions of the bottle since these solutions may contain a very powerful acid. Alloys receiving an *A* or *B* corrosion rating in **table 1** require no further treatment to prevent destructive corrosion for several years of use. A coating of clear lacquer or acrylic spray may be used to protect the porous metal surface from stains and cosmetic corrosion.

conclusion

I don't expect anyone to start manufacturing chassis boxes at home; the process is far too time consuming. The main point in having a simple bending tool is to enable you to fabricate a special part in a size and shape not available commercially, with a professional, high-quality appearance.

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1. *Product and Data Catalog*, Reynolds Aluminum Supply Company, free to Reynolds customers at Reynolds distributors.
2. *Metals Handbook*, 8th Edition, Volume 1 (of II volumes), American Society for Metals, Metals Park, Ohio, 1977.

ham radio

a regulated screen grid power supply

An easy-to-build solid-state design using Motorola's new TMOS power-FET technology

For about the last 20 years, the most common type of amplifier used by hams has been the grounded-grid, class AB2 linear, sometimes erroneously referred to as "Class B." In Class B operation, no quiescent plate current is drawn; Class AB2, on the other hand, has quiescent plate current. Both classes run grid current when a signal is present. If there were no idling current in a grounded-grid amplifier, it wouldn't be linear. There is no such thing as a single-ended "grounded-grid Class B linear."

Every linear amplifier type has some advantages and some tradeoffs. One advantage of a grounded-grid amplifier is that it supposedly does not need to be neutralized. This is probably true for 813 tubes, which wouldn't take off and oscillate on a VHF parasitic frequency even if you wanted them to. An 813 doesn't even like to work at 29 MHz. Tubes such as the 3-500Z are designed to amplify well, or oscillate, at 100 MHz.

During the last ten years I have been involved in more than 30 cases of amplifier repair. More than 80% of these failures involved damage from parasitic oscillation in grounded grid amplifiers that supposedly do not need to be neutralized. A stable grounded grid amplifier design is not as easy as the popular literature on the subject indicates. This subject is covered in detail in the April 1986 issue of this magazine.

Another advantage of the grounded-grid configuration is that no screen or control grid bias supplies are needed. Control grid bias supplies are simple for Class AB-1 grid-driven amplifiers, since, typically, no current flows. Screen grid supplies are a different matter because they must also furnish current — sometimes over 100 mA. One disadvantage in grounded-grid operation is that separate tuned inputs must be constructed for each band in use.

The grid-driven linear offers certain advantages. A separate tuned input is not required for each band, and up to approximately 23 dB of amplification can be obtained instead of the typical 12 dB gain from grounded-grid. This means you can drive a legal-limit amplifier with a 10-watt QRP transceiver. To do this you'd have to neutralize the amplifier; this would complicate the grid input circuit because the easy, 50-ohm grid swamping resistor circuit would not work. One disadvantage, however, is that you'd have to build a high-voltage screen supply. For a 4-1000A, a screen voltage of 1000 volts is required for Class AB1 linear amplifier service. With 6000 volts on the plate, this tube will deliver 1500 watts output. Another good tetrode for ham service is the 4CX1500A, which requires up to 750 volts for the screen grid.

The choice between grounded-grid and grid-driven Class AB1 would have to go to grounded-grid up until the World Administrative Radio Conference of 1979; until that time, amplifiers were in use on only about five Amateur Radio bands below 30 MHz, and building five tuned inputs for the common grounded-grid Class AB1 amplifier posed no problem. The advantages of using grid-driven Class AB1 amplifiers were not yet apparent, especially in light of the grief of building a regulated high-voltage screen supply. Most people would rather build the separate tuned inputs for the simpler grounded-grid triode circuit than use the inherently broad-band Class AB1 input circuit.

Today amplifiers are used on eight Amateur bands below 30 MHz, and can be used with reduced output on the 30-meter band. In a while we'll have nine bands on which amplifiers can be used below 30 MHz. Building nine separate Pi-networks requires 27 components

By Richard Measures, AG6K, 6455 LaCumbre Road, Somis, California 93066

and a rare nine-position bandswitch; in comparison, the Class AB1 input circuit begins to look very simple. Perhaps building a screen supply isn't really that much of a problem after all. . . .

For the Amateur who is also a MARS member, the new general-coverage transceivers are a blessing. A general coverage amplifer would be a nice accessory for a general coverage transceiver.

to regulate or not

According to Eimac, "The power output from a tetrode or pentode is very sensitive to screen voltage. For this reason, any application requiring a high degree of linearity through the amplifier requires a well-regulated screen supply."[1]

For some small tetrodes, it's possible to use gas voltage regulator tubes. For larger tetrodes with higher than about 25 mA of screen current the VR tube isn't practical; a better choice is a conventional series pass regulator circuit, which offers the added benefit of adjustable screen voltage, to either control power output or to set the idling plate current to the desired value, a nice feature for those times when 100 watts isn't enough, but 1500 watts is too much. If a roller-coil is used for the plate tank inductor, a general coverage 1.8 to 30 MHz amplifier is yours. The only band-switching operation needed is that of adding loading capacitor padders for frequencies below about 5 MHz.

With the common grounded-grid amplifier, reducing power is not easy; in a Class AB1 amplifier, it's simple. Series pass transistor regulator circuits are used in 12- to 14-volt power supplies for Amateur transceivers and other low-voltage applications. Screen supplies for Class AB1 amplifiers typically must provide between 500 and 1500 volts. Until recently, a regulated supply for these voltages required tube circuits because transistors couldn't handle the required voltages and currents.

Bipolar transistors that would handle high voltages, at low current, for TV applications have been around for some time. The Motorola BU207 was one example. It could withstand either high voltage (up to 1300 volts) or high current — but not both at the same time, because of secondary breakdown effects: a 1500-volt, 8 ampere bipolar may be able to hadle only 0.2 amps at 200 volts. Your 100-watt transistor is now a 40-watt transistor. Three transistors are required for 100 watts dissipation. High voltage bipolar transistors have another trade-off: low current gain. So you need a driver transistor to provide high base current. It was possible to use bipolars in high-voltage regulated supplies, but it took a bank of them to form the series pass element. (The first high-voltage regulated supply I built used six BU207 bipolars to produce 1500 volts at 0.2 amps. It worked, but it wasn't as simple as I'd have liked it to be.)

There has been progress, however. Just as FET RF power transistors are taking over jobs from bipolars, the high-voltage FET is now available for use in high-voltage applications such as regulated supplies. Motorola is using this new technology in its TMOS power FET. The TMOS offers high gain and virtually no secondary breakdown problems that would reduce the dissipation rating.

A TMOS FET is similar to a triode electron tube; the more positive the control element, the more current flows in the load. Tubes have some nasty disadvantages: they lose linearity when the control grid goes positive and start to draw lots of grid current at the same time. TMOS devices don't exhibit this effect. The gain of a triode is rated in micromhos — or more recently, in microSiemens. A triode with a transconductance of 25,000 micromhos is considered a "hot" tube.

The TMOS device I used to construct the power supply for this article has a transconductance of about 500,000 microSiemens in its linear range. As you may have guessed, the device drives so easily it can be a problem to keep it *from* taking off unless you swamp the control gate with a resistor and a capacitor.

design considerations

In a typical 13.8-volt supply the filter capacitor is charged to about 22 volts. If a short circuit were placed across the output, 22 volts would appear across the pass transistor — until the fuse in the primary burned out. In a 1500-volt supply the filter capacitor can be charged to 1900 volts in the no-load condition. In this case a short would place 1900 volts on the pass transistor. Twenty-two volts isn't a problem for a typical 40-volt bipolar, so no protection is needed; 1900 volts, however, is a completely different matter. To protect the high-voltage power supply series pass element from short circuits, some kind of pass element crowbar is required. I used an SCR whose gate is fired by a string of zener diodes when the voltage gets into to danger region above 520 volts difference across the pass element. Since it takes about 2 microseconds to fire the SCR, an RC delay circuit is used (R3C1) to protect the pass element while the SCR is reaching threshold. R2 limits the current through the SCR to less than its rated 160 peak amps. R2 is 10 ohms for up to 1000 volts output and 20 ohms for up to 2000 volts output.

The voltage rating of the pass element needs to be slightly greater than the difference between no-load voltage across the filter capacitor and the desired regulated output voltage. Naturally, the power supply must provide enough voltage at full load or regulation is not going to occur. In addition, the product of current and voltage across the pass element must not exceed its dissipation rating.

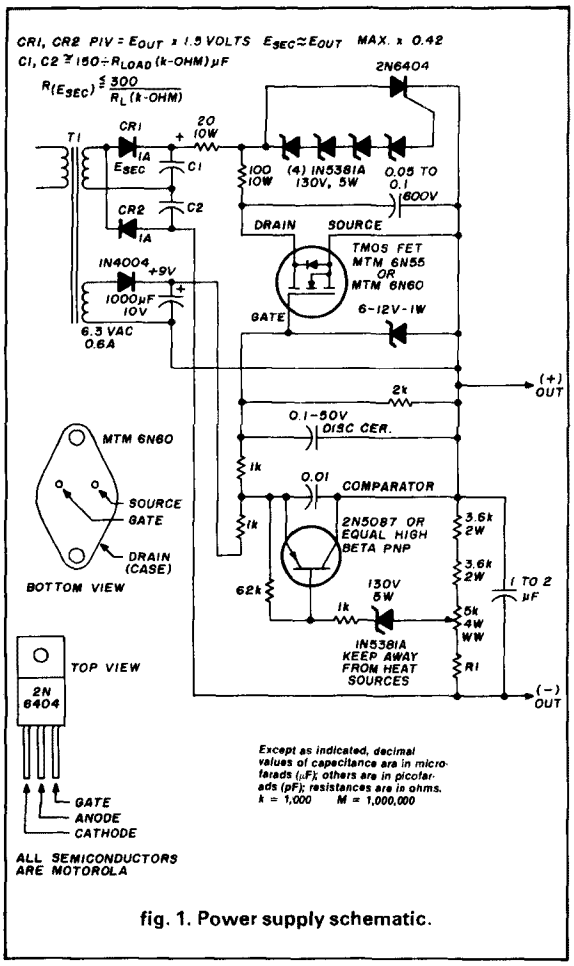


fig. 1. Power supply schematic.

Most capacitor filter power supplies see a voltage drop of about 15 per cent from no-load to full load. This means that the pass element needs to withstand only about 15 percent of the output voltage plus a small safety factor. A 600-volt pass element can easily regulate a 3000-volt supply. The highest rated voltage available in the Motorola TMOS series is 1000 volts, which means you could build a regulated 5000-volt supply with an output current of 2 amps if you needed one. To achieve the current requirement it would be necessary to parallel some TMOS devices and connect equalizing resistors of a few ohms in each source lead. The comparator circuit, Q1, is simplicity itself, since no driver transistor is required for the TMOS pass device.

circuit operation

The positive output terminal is the common reference point, or circuit common. When the output voltage rises to a high enough negative value V_z begins to conduct, causing Q1 to conduct. This increases the voltage drop across R4, causing the gate voltage of Q2 to be less positive, which causes Q2 to conduct less. If Q2 is conducting less, then the output voltage begins to fall because of the increased voltage drop across the pass element. As the output voltage falls, V_z conducts less, causing Q1 to conduct less. This causes Q2's gate to go more positive. Q2 conducts more heavily, causing the output voltage to rise again. The process repeats and regulation results.

The threshold voltage for turn-on/turn-off is determined by the voltage division ratio of the R1, R5, R6 voltage divider and the V_z zener voltage. R6 is adjusted to the desired output voltage as long as not much more than 500 volts is dropped across the pass device. Setting the output voltage too low fires the pass-protect crowbar, causing the output voltage to rise to the voltage on the power supply filter capacitor less about 1 volt. At this point you need to turn the power off and on, unlocking the crowbar, and then make a mark on the R6 scale to indicate how low you can go. Starting the power supply with a heavy load in place will also fire the crowbar. This isn't a problem in screen supplies because the amplifier won't be operated until the final amplifier tube's filament comes up to temperature. This is about 1 second for a 4CX1500A. If you want to be able to start with a heavy load connected, you'll need to add one or more zener diodes from the drain to the gate of Q2 to start conduction when the voltage drop is about 20 volts less than the crowbar trigger voltage.

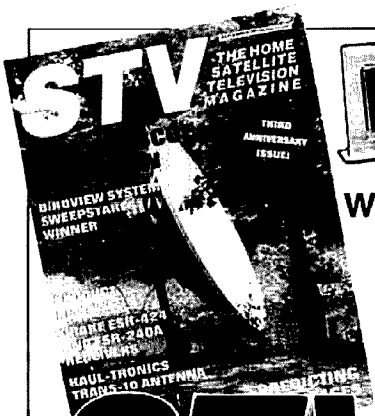
performance

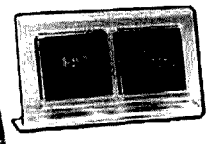
The load regulation is more than adequate for this application. With the output set for 900 volts on my prototype supply, the voltage drops to 896 volts with a 225 mA load. Reverse screen current up to about 20 mA is no problem. Beyond 20 mA reverse current, you'll need to add an appropriate bleeder resistor across the output terminals.

This screen supply can provide conventional positive output, or, negative output for grounded screen Class AB1 grid-driven amplifiers, in which the positive output terminal is grounded and the screen negative voltage is connected to the cathode or filament transformer center-tap. This configuration is used in many commercial short-wave transmitters because there's no need for a screen bypass capacitor and the unwanted added inductance associated with it.

choosing the right fuse

It's important to fuse this supply in order to protect the screen of your amplifier tube from excessive dissipation, which will never occur unless you tune up under light load conditions, as might occur, for example, if your antenna were not connected. Light loading allows the instantaneous plate voltage to fall to a level that fails to attract the electrons coming from





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the cathode. These electrons will then find the relatively positive screen very attractive, causing an unhappily large flow of current in the screen grid. The screen grid will suffer a melt-down unless something is done quickly. An ordinary fuse will perform this duty faithfully if the right ampere rating is used.

With a capacitor filter supply, which normally has high peak currents in the primary of the power transformer, a fuse that is about 1.5 times larger than the screen dissipation rating would indicate is adequate. For example, with a 120-VAC primary, a 1-amp fuse would allow 120 watts to flow. This would be adequate for a screen grid rated at 80 watts maximum dissipation. The screen supply transformer should be connected to the same step-start source used for the plate transformer so that the regular fuse won't burn out at turn-on. A one-third duty cycle tuning pulser will save fuses if you make an error during tune-up.

choosing the transformer

When you're selecting a power transformer for a screen supply, "heavy duty" is *out*. You don't want a screen supply that can do double duty as an arc-welder. A screen supply transformer should have about the same CCS current rating as the typical screen current on the tube's spec sheet. The transformer selected should also have a filament winding to provide the voltage to operate the comparator circuit, as shown in **fig. 1**. The current drawn by this circuit is about 6 mA — so any 6.3-volt or 12.6-VCT winding will do the job. Other transistors can be used in the comparator as long as they are high Beta and PNP.

Motorola TMOS power FETs have many other interesting uses besides series pass regulators. The MTM 6N60 I used in this article will make a nice shunt regulator at up to 600 volts at 100 watts dissipation — if you keep it cool. If that's not enough, the MTM 6N60 has a big brother, the MTE 40N60, which is a 40 amp, 600-volt device.

I believe we're going to see a lot more TMOS devices in Amateur Radio equipment. Military transceivers already use TMOS output amplifiers. How about 300 watts output on 2 meters with 5 watts of drive? That's a pair of Motorola MRF 150s, which will do the same job on 160 meters. TMOS RF amplifiers produce less intermodulation distortion than conventional bipolar RF amplifiers. Their use could represent a pleasant improvement on the ham bands. With RF-negative feedback it is possible to make a solid-state TMOS SSB signal as clean as some of the cleanest tube-type equipment such as the TS830S.

references

Care and Feeding of Power Grid Tubes, EIMAC Division, Varian Inc., 301 Industrial Way, San Carlos, California 94070.

ham radio

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direct currents reduce core permeability

BASIC program shows how to compensate for this effect

Applications involving coils that carry DC are common. Perhaps not so common, however, is the understanding that inductance determination requires a knowledge of the anticipated DC current. A DC bias reduces the inductance value and must be taken into consideration for accurate results.

Although this toroidal-inductor design program was written for Microsoft™ BASIC and tested on a Commodore Pet Computer with 32k of memory, no special machine-dependent functions were used; therefore, it can be used on virtually any home computer with little or no modification. Remark statements are liberally provided throughout the program listing to identify operations.

design procedure

The design sequence of calculations for each wire size is as follows:

- Calculation of the number of turns (N) assuming a winding factor of 0.525.
- Calculation of DC resistance (RDC).
- Calculation of copper weight.
- Calculation of maximum DC current.
- Calculation of inductance (mH).
- Calculation of wound outside, inside diameter, and height of wire and core.
- Calculation of temperature rise in No. 30 AWG wire considering copper losses only.

A complete enamel wire table is included in the program for wire sizes No. 12 AWG through No. 46 AWG. Coils are designed to have a winding factor of 52.5 percent maximum.

Selection of a core size and core material for a desired inductance can be determined by doing a set of iterative calculations using vendor curves (**fig. 1**) showing percent permeability reduction versus DC. The magnetizing force, H, (in oersteds) is:

$$H = 0.4\pi NI/P \qquad (1)$$

where N = number of turns
I = current in amperes
P = magnetic path length in centimeters

In **fig. 1**, a certain percent permeability reduction is initially assumed for material of the selected permeability. This establishes the maximum magnetizing force, H, in oersteds, which must not be exceeded if the application's determined requirements are to be met. AC flux density (BAC) is assumed to be small.

The design program uses this information and other core data from the selected vendor data sheet to compute a table of inductor designs.

Toroidal Molypermalloy Powder (MPP) cores are a useful alternative for gapped cores in applications where there is a direct current component which tends to saturate the core. MPP cores are well suited for low inductances, below 2 mH. Advantages of using MPP cores are largely in elimination of assembly and labor costs for handling two core halves, fitting gap spacers and mechanical parts.

One vendor has published a core and permeability selection procedure that simplifies the choice of a core. Because selection of the core to avoid saturation is a most significant choice in inductor design, a modified version of this procedure, courtesy of Magnetics, Inc.,* is used here as a part of the computer design program user instructions.

program user information

Only two parameters for each (design) application must be known: inductance required under DC bias conditions and the amount of DC current.

Calculate the product LI^2 where:
L = inductance required with DC bias (millihenries)
I = DC current (amperes).

Locate the LI^2 value on the DC bias core selector chart (**fig. 2**). Follow this coordinate to the intersec-

By Walter Kunde, K5BVM, Route 3, Box 346, Quinlan, Texas 75474

*Magnetics, Inc., P.O. Box 391, Butler, Pennsylvania 16003.

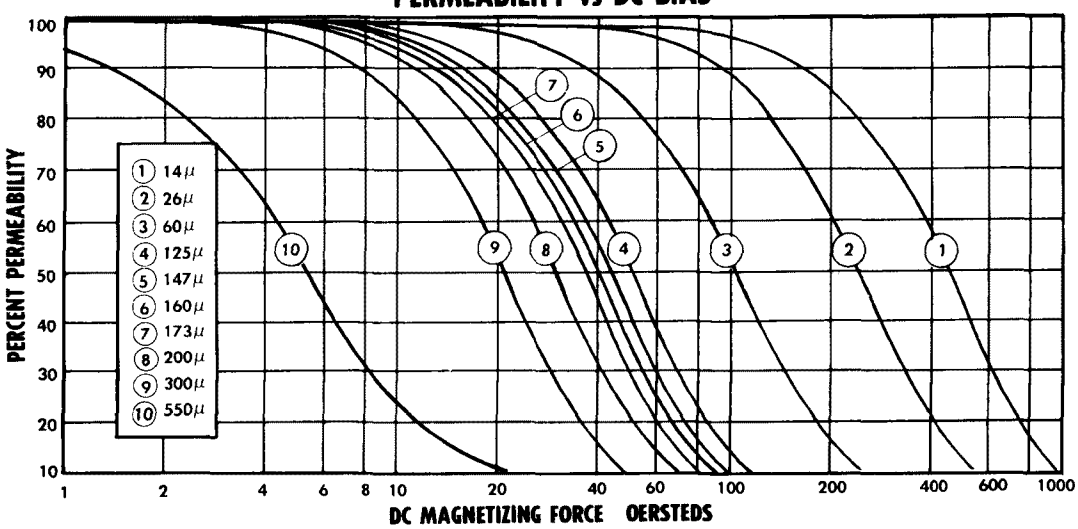


fig. 1. Permeability versus DC bias.

tion with the first core size that lies within the family of solid permeability lines. This core size is the smallest that can be used. (In **fig. 2**, small core sizes are at the bottom and large sizes at the top.)

Any solid permeability line that passes through the intersection point of the LI^2 and core size coordinates or crosses the LI^2 coordinate below this intersection point may be used.

The choice of permeability can be based on Q requirements at the operating frequency of the application. Use the Q curves in **fig. 3** for this selection. If Q is not a consideration, use the highest permeability indicated. This choice will yield the lowest winding factor.

The design application may call for a Q requirement at a higher frequency than the core/permeability combinations indicated in **steps 2 and 3**. By following the LI^2 coordinate to lower permeabilities and larger core sizes, an optimum choice for this type design may be made. For a given permeability, always use the core size just above the permeability line.

Inductance, core size and permeability are now known. The remaining core data is obtained from the selected core data sheet, see **fig. 4**. Use core dimensions which include the core finish.

Load the program and respond to input commands when requested.

Inputs include the following:

- core outside diameter (in inches)
- core inside diameter (in inches)
- core height (in inches)
- core permeability
- core mH/1000 turns
- magnetic path (in inches)
- window area (in circular mils) — CMIL* $\times 10^6$
- MLT (mean length of turn) — full winding (in inches)
- HDC (Peak) Maximum
- percent Nom. Perm @HDC (Peak) Maximum

Upon completion of the last input, "**CALCULATING**" will appear on the monitor screen as the program calculates all values. Results are displayed as printer outputs; see **figs. 5 and 6**.

core selection example

Choose a core meeting the following requirements:

- minimum inductance with DC bias, 25 μH
- DC current, 2 amps.
- operating frequency, 8 kHz.
- optimum Q is required.

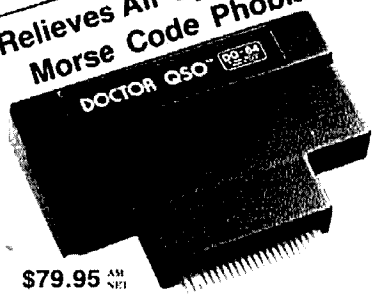
The product of $LI^2 = 0.1$. Following this coordinate, the first core size encountered that falls within the solid line permeability family is the 55050 size.

The intersection point of the 0.1 coordinate and the 55050 core size coordinate falls between the 60μ curve and the 125μ curve. Only those permeability lines intersecting the LI^2 coordinate below the core intersection are usable; either 125μ or 160μ may be used.

A review of the Q curves (**fig. 3**) indicates that a 55050 core in 160μ peaks in the neighborhood of 8 kHz, whereas this size in 125μ peaks near 20 kHz. For optimum Q at 8 kHz, 160μ is chosen, core 55048. From **fig. 1**, determine magnetizing force H for the percent-

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DC BIAS CORE SELECTOR CHART

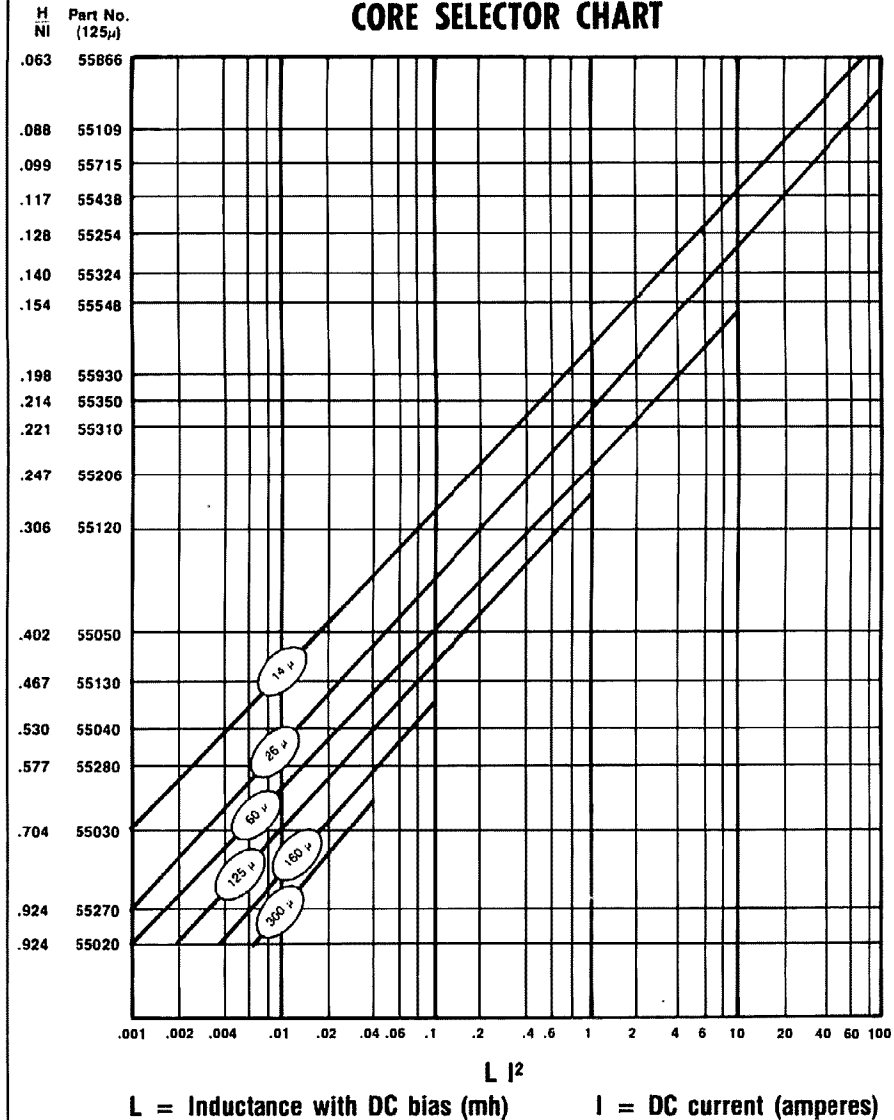
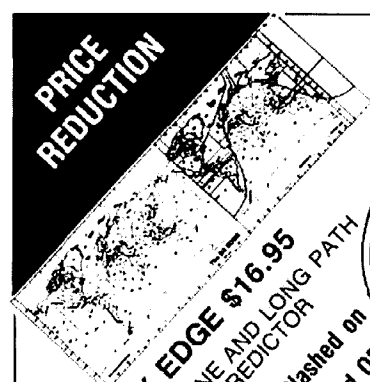
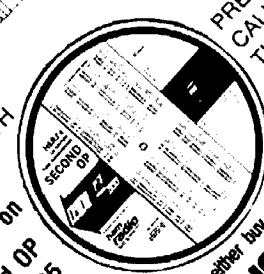


fig. 2. DC bias core selector chart.

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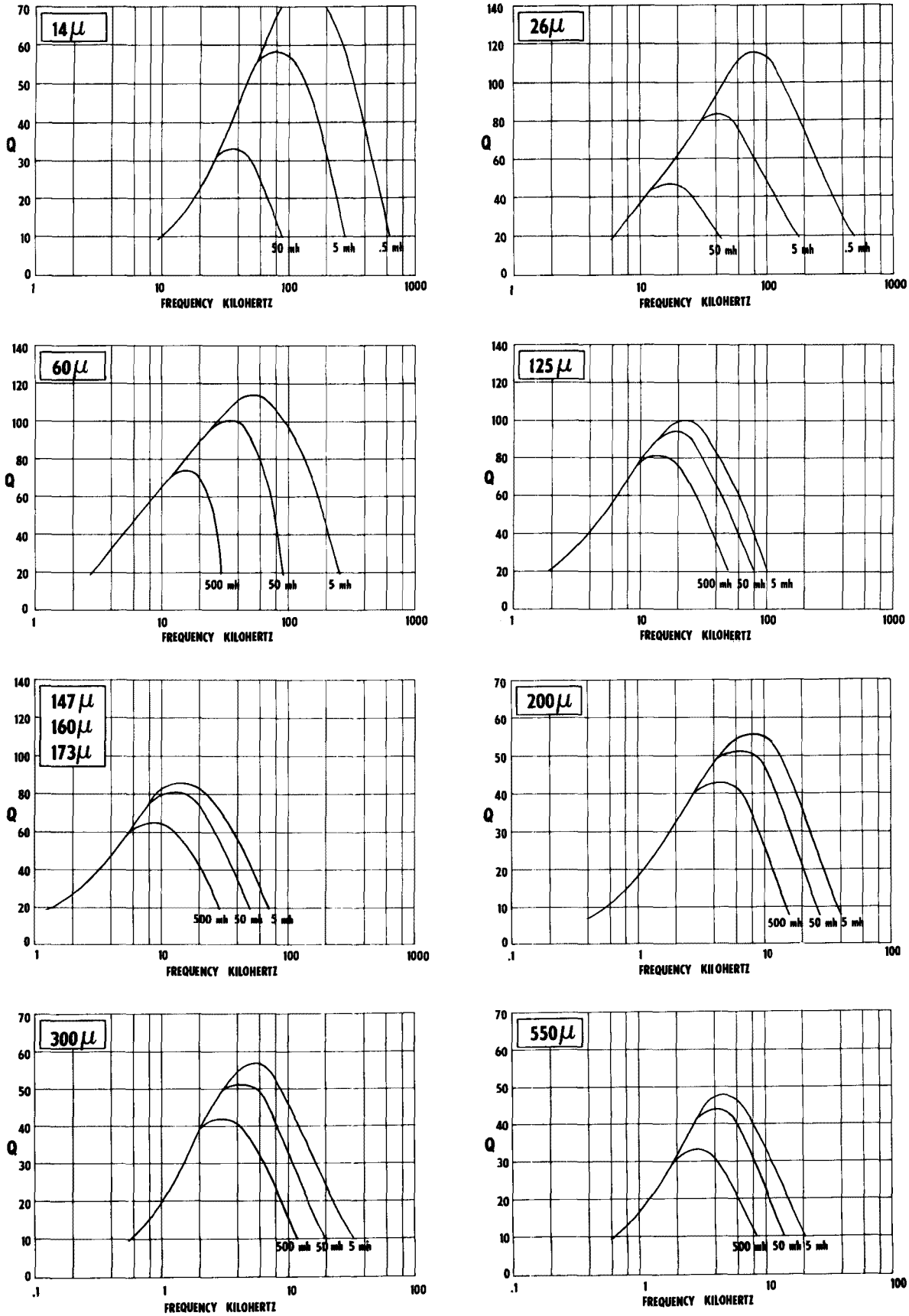


fig. 3. Using Q curves helps determine permeability.

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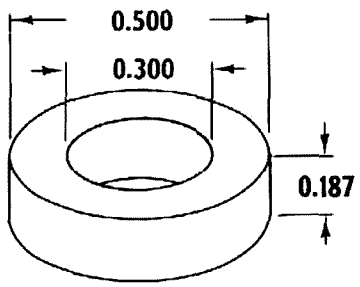
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Intl TELEX 6502881526 WUI

✓ 145



WINDOW AREA	75,600 cir. mils	
CROSS SECTION	0.01767 in^2	0.1140 cm^2
PATH LENGTH	1.229 in	3.12 cm
WEIGHT	0.11 oz (.0089 lb.)	3.1 gm

WINDING TURN LENGTH

WINDING FACTOR	LENGTH/TURN	
100% (UNITY)	0.0815 ft	2.49 cm
60%	0.0721 ft	2.20 cm
40%	0.0629 ft	1.920 cm
20%	0.0589 ft	1.797 cm
0%	0.0574 ft	1.751 cm

CORE DIMENSIONS AFTER FINISH

OD (MAX.)	0.530 in	13.46 mm
ID (MIN.)	0.275 in	6.99 mm
HT (MAX.)	0.217 in	5.51 mm

WOUND COIL DIMENSIONS

UNITY WINDING FACTOR

OD (MAX.)	0.717 in	18.2 mm
HT (MAX.)	0.451 in	11.5 mm

MAGNETIC INFORMATION

PART NO.	PERM μ	INDUCTANCE @ 1000 TURNS MH ±8%	NOMINAL DC RESISTANCE OHMS/MH**	FINISHES AND STABILIZATIONS*	GRADING STATUS 2% BANDS	B/NI GAUSS PER AMP TURN
55053-	14	6.4	2.67	A2	*	5.64 (@ 1500 gauss)
55052-	26	12	1.42	A2	*	10.5 (@ 1500 gauss)
55051-	60	27	0.633	ALL	*	24.2 (@ 1500 gauss)
55050-	125	56	0.305	ALL	YES	50.4 (@ 1500 gauss)
55049-	147	67	0.255	ALL	YES	59.2 (@ 1500 gauss)
55048-	160	72	0.237	ALL	YES	64.5 (@ 1500 gauss)
55044-	173	79	0.216	ALL	YES	69.7 (@ 1500 gauss)
55047-	200	90	0.189	ALL	YES	80.6 (@ 600 gauss)
55045-	300	134	0.127	A2 and L6	YES	121 (@ 300 gauss)
55046-	550	255	0.067	A2	YES	222 (@ 50 gauss)

WINDING INFORMATION

*FOR UNITY WINDING FACTOR***

AWG WIRE SIZE	TURNS	Rdc OHMS	AWG WIRE SIZE	TURNS	Rdc OHMS
16	25	0.00826	31	646	6.90
17	31	0.0129	32	788	10.40
18	39	0.0203	33	977	16.40
19	48	0.0318	34	1,244	26.5
20	61	0.0500	35	1,543	41.6
21	76	0.0789	36	1,905	64.4
22	93	0.1233	37	2,327	97.1
23	116	0.1925	38	2,909	153.6
24	144	0.302	39	3,744	258
25	178	0.470	40	4,727	414
26	222	0.743	41	5,817	626
27	280	1.173	42	7,414	1,003
28	344	1.829	43	9,003	1,570
29	420	2.81	44	10,360	2,186
30	525	4.45			

fig. 4. Core data.

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*** TORCHK1 PROGRAM ***

CORE O.D.: .53

160 PERM.

WIRE SIZE	FULL COIL TURNS	DC RES OHMS	CU WT	I PEAK AMPS	INDUC MHY
12	5.4	0	.01	9.19	1.6E-03
13	6.7	0	.01	7.41	2.5E-03
14	8.4	0	.01	5.91	4E-03
15	10.5	0	.01	4.72	6.3E-03
16	13.2	0	.01	3.76	.01
17	16.3	0	.01	3.04	.0153
18	20.5	0	.01	2.42	.0242
19	25.4	.01	.01	1.95	.0371
20	31.8	.02	.01	1.56	.0582
21	39.4	.03	.01	1.26	.0894
22	49.1	.05	.01	1.01	.1388
23	61	.08	.01	.81	.2143
24	75.7	.14	.01	.65	.33
25	93.6	.21	.01	.53	.5046
26	116	.34	.01	.42	.775
27	145.9	.54	.01	.34	1.2261
28	181.2	.85	.01	.27	1.8912
29	220.4	1.29	.01	.22	2.7979
30	275.6	2.06	.01	.18	4.375
31	339.2	3.2	.01	.14	6.6272
32	413.4	4.82	.01	.12	9.8438
33	512.7	7.61	.01	.09	15.1408
34	652.7	12.28	.01	.07	24.5385
35	809.9	19.32	.01	.06	37.782
36	999.7	29.9	.01	.04	57.5654
37	1221.2	45.07	.01	.04	85.9005
38	1526.5	71.3	.01	.03	134.2196
39	1964.8	119.96	.01	.02	222.3612
40	2480.6	193.11	.01	.02	354.4344
41	3053	290.49	.01	.01	536.8785
42	3891.1	465.59	.01	.01	872.1019
43	4724.9	728.86	.01	.01	1285.9015
44	5436.9	1015.04	.01	0	1702.6491
45	7488.6	1807.25	.01	0	3230.1578
46	9020.4	2735.47	0	0	4686.7746

****

fig. 6. Table of inductor designs.

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- Standard DB25 for RS232 connection
- Simple radio hookup
- Radio modem w/built-in counter for calibration
- Low power CMOS option
- Tuning indicator socket for HF & satellite work
- Modem disconnect for future options
- Lithium battery backup for RAM



Hardware
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Documentation | by TAPR

Model TNC2A Kit NMOS $154.95
Model TNC2A Kit CMOS $169.95

SPECIFICATIONS

CPU	Z80A microprocessor
Clock	7.4576 Mhz standard, 4.9152 Mhz available
Memory	32K EPROM, 16K RAM standard
HDLC	Packets are controlled by hardware for maximum performance, permitting full duplex operation
Modem	1200 baud, Bell 202 compatible standard, easily configured for 300 baud/200 Hz shift for HF use
Serial	Computer/Terminal port is industry standard RS 232 C compatible for use with most equipment
Radio	Watchdog timer for channel protection transmits audio levels adjustable for nearly any radio. Wide dynamic range demodulator. Channel busy input (RF DCD) to inhibit packet transmissions on a shared channel
LEDs	Power - tells you when power is applied Status - tells you when you have unacknowledged traffic on your buffers Connect - tells you when you are in the error free mode DCD - tells you when your TNC2A senses other activity on the channel PTT - tells you when your TNC2A keys the transmitter
Power	+10 to +15 VDC CMOS 110 ma NMOS 260 ma Typical

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8241	1500B	RG59/U Poly 96%	13.00	.15
8287	1130B	RG213/U Poly 96%	53.00	.59
9269	1600B	RG62A/U Poly 96%	15.00	.17
8216	1450B	RG174/U Poly 96%	12.00	.14
9913	1180	Low Loss 50 Ohm	46.00	.58
		OTHER QUALITY CABLES		
Nemal No.		Description	Per 100 ft.	Per ft.
1110		RG8X 95% Shield (mini 8)	15.00	.17
1130		RG213/U Mil Spec. 96% Shield	34.00	.36
1140		RG214/U Mil Spec. - Silver	155.00	1.65
1705		RG142B/U Teflon - Silver	140.00	1.50
1310		RG217/U 5/6" 50 Ohm Dbl. Shld.	80.00	.85
1470		RG223/U Mil Spec. - Silver	80.00	.85
		ROTOR CABLE — 8 COND.		
8C1822		2-18 Ga., 6-22 Ga.	19.00	.21
8C1620		2-16 Ga., 6-20 Ga. Heavy Duty	34.00	.36
		HARDLINE — 1/2"		
FXA12		Smooth Alum. w/black jacket	79.00	.89
FLC12		Corrug'd. Copper (EQ. Heliax LDF)	159.00	1.69
		CONNECTORS — MADE IN U.S.A.		
NE720		Type N for Belden 9913		4.75
PL259		Standard Plug for RG8, 213		.65
PL259AM		Amphenol PL259		.89
PL259TS		PL259 Teflon/Silver		1.59
UG21D		Type N for RG8, 213, 214		3.00
UG175		Adapter for RG58		.22



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glossary of terms

Winding factor is the ratio of the total area of copper wire in the center hole of a toroid to the window area of the toroid.

Window Area, CMIL • 1E6 circular. The unit "Circular MIL" is the area of a circle 1 mil in diameter. The number of circular mils in a square inch is, therefore $\frac{4}{\pi} \times 10^6$, or 1,273,237. Since 1 mil is the thousandth part of an inch; there are 1,000,000 square mils in a square inch. Thus the area of a cross-section expressed in circular mils is always bigger than the true area expressed in square mils, because the unit area to which the name "Circular Mil" has been given is smaller than the square mil.

Heavy Enamel is a magnet wire having an extra heavy film of enamel insulation (i.e., double coated). Its abrasion resistance is somewhat greater than regular enamel (i.e., single coated) magnet wire.

MLT — full winding. MLT is the mean length of a single turn of wire. This will vary considerably between a single layer of wire and a core *full* of wire. The values of MLT are listed on each core size data sheet (see **fig. 4**).

HDC (peak) Max. (core bias). DC current = peak current. If current is not direct continuous, then the peak (momentary) value is used to determine core operating conditions.

$$H = \frac{0.4\pi \cdot turns \cdot peak\ magnetizing\ current}{mean\ magnetic\ path}$$

or $H = \frac{0.4\pi \cdot NI}{\ell}$ *Oersteds*

bias range Oersteds	recommended permeability
below 6	500
6 to 22	300
22 to 35	200
22 to 35	173
22 to 35	160
35 to 70	147
40 to 140	125
70 to 150	60
150 to 330	26
above 330	14

core O.D.	0.53
core ID	0.275
core height	0.217
core permeability	160
MHY/1000T (mH)	72
magnetic path	1.229
window area	0.0756
mean turn	0.865
HDC peak maximum	20
percentage MU @ HDC peak maximum	0.8
wound OD	0.56
wound ID	0.18
wound height	0.31

Note: Temperature rise for No. 30 AWG (copper loss only) is approximately 10.53 degrees C.

fig. 5. Input summary and wound inductor dimensions for a powdered iron core that is carrying maximum peak current.

fig. 7. Basic program compensates for core permeability.

```
10 REM   ********************************************************************
20 REM          POWDERED IRON CORES
30 REM      PROGRAM GENERATES INDUCTOR
40 REM         DESIGNS CARRYING MAX.
50 REM        VALUES OF PEAK CURRENT.
51 REM            FEBRUARY23,1985
52 REM            WALTER KUNDE,P.E.
59 ' .
60 REM          TORCHK1 PROGRAM NOTES
70 '
80 REM    1.USES WIRE SIZES 12AWG TO 46AWG.
120 REM   2.USE DATA FROM MAGNETICS, BUTLER, PA CATALOG MPP-383S.
121 REM   3.CALCULATED I PEAK CONSIDERS CORE HDC MAX. ONLY.
122 REM     CURRENT CAPACITY OF WIRE IS SEPARATE CONSIDERATION AND INVOLVES THE
123 REM     DUTY CYCLE ALSO.
130 '
140 REM   ********************************************************************
150 '
170 DIM L(35),N(35),CMHEVV(35),PDS(35),IDADE(35)
180 DIM OHMS(35),FEET(35),ROC(35),CWT(35),APKMAX(35)
190 '
192 REM   CMHEVV DATA (INCLUDES WIRE INSULATION)
193 '
200 DATA 73.18E+2,58.52E+2,46.79E+2,37.58E+2,30.03E+2,24.21E+2,19.36E+2
210 DATA 15.68E+2,12.46E+2,10.05E+2,807.,650.,524.,424.,342.,272.,219.,180.
220 DATA 144.,117.,96.8,77.4,60.8,49.8,39.7,32.5,26.8,20.2,16.8,13.8,10.2,8.4
230 DATA 7.3,5.3,4.4
240 '
250 REM   OHMS/FT DATA
260 '
270 DATA 1.59E-3,2.00E-3,2.52E-3,3.18E-3,4.02E-3,5.05E-3,6.39E-3,8.05E-3
280 DATA 10.1E-3,12.8E-3,16.2E-3,20.3E-3,25.7E-3,32.4E-3,41.0E-3,51.4E-3
290 DATA 65.3E-3,81.2E-3,104.E-3,131.E-3,162.E-3,206.E-3,261.E-3,331.E-3
300 DATA 415.E-3,512.E-3,646.E-3,847.E-3,108.E-2,132.E-2,166.E-2,214.E-2
310 DATA 259.E-2,3.340,4.207
320 '
330 REM   POUNDS/FT DATA
340 '
350 DATA 20.1E-3,15.9E-3,12.6E-3,10.0E-3,7.95E-3,6.33E-3,5.03E-3,3.99E-3
360 DATA 3.18E-3,2.53E-3,2.00E-3,1.60E-3,1.26E-3,1.00E-3,0.794E-3,0.634E-3
370 DATA 0.502E-3,0.405E-3,0.319E-3,0.253E-3,0.205E-3,0.162E-3,0.127E-3
380 DATA 0.101E-3,0.805E-4,0.655E-4,0.518E-4,0.397E-4,0.312E-4,0.254E-4
390 DATA 0.203E-4,0.157E-4,0.131E-4,9.40E-6,7.50E-6
400 '
401 '
410 REM   IDADE DATA
420 '
430 DATA 12,13,14,15,16,17,18,19,20,21,22,23,24,25,26,27,28,29,30,31,32,33,34
440 DATA 35,36,37,38,39,40,41,42,43,44,45,46
450 '
470 '
471 PRINT "[2J       *** TORCHK1   INPUTS ***"
472 PRINT
480 INPUT "CORE OUTSIDE DIA.IN.";OD
481 PRINT
490 INPUT "CORE INSIDE DIA.IN.";ID
491 PRINT
500 INPUT "CORE HEIGHT IN.";HT
501 PRINT
510 INPUT "CORE PERMEABILITY";U
511 PRINT
520 INPUT "CORE MHY/1000 TURNS";MHY
521 PRINT
530 INPUT "MAGNETIC PATH IN.";MMP
531 PRINT
540 INPUT "WINDOW AREA CMIL*1E6 ";AW
541 PRINT
550 INPUT "MLT-FULL WINDING-IN.";MLT
551 PRINT
560 INPUT "HDC (PEAK) MAX.";MPK
561 PRINT
570 INPUT "% NOM PERM @ HDC (PK) MAX.";P
571 PRINT
580 PRINT "          ** CALCULATING ** "
590 '
600 REM **** CALCULATE CHARACTERISTICS OF FULL COIL FOR EACH WIRE SIZE ****
610 '
620 '
640 '   FOR J=1 TO 35
650 '   READ CMHEVV (J)
660 '   N(J)=INT((10)*((AW*.525*1000000)/(CMHEVV(J))))/(10)
670 '   NEXT J
679 '
690 '
700 REM 0.525=.70(FILL FACTOR)*.75(AVAILABLE AREA) = WINDING FACTOR
710 REM NOW CALCULATE DC RESISTANCE OF WINDINGS @ 25 DEGREES C.
720 '
740 '   FOR J =1 TO 35
750 '   READ OHMS(J)
760 '   FEET(J)=INT(0.5+(10^2)*(MLT*N(J)/12))/(10^2)
770 '   ROC(J)=INT((10^2)*(FEET(J)*OHMS(J)))/(10^2)
780 '   NEXT J
790 '
800 REM  NOW CALCULATE COPPER WEIGHT OF WINDINGS.
820 '   FOR J =1 TO 35
830 '   READ PDS(J)
840 '   CWT(J)=INT(0.5+(10^2)*(FEET(J)*PDS(J)))/(10^2)
850 '   NEXT J
860 '
870 REM *** NOW CALCULATE PEAK CURRENT IN AMPERES ***
880 '
900 '   FOR J=1TO35
910 '   APKMAX(J)=INT((10^2)*(MPK*MMP*2.82/N(J)))/(10^2)
920 '   NEXT J
940 '
941 '   FOR J=1 TO 35
942 '   READ IDADE(J)
943 '   NEXT J
944 '
950 REM *** NOW CALCULATE INDUCTANCE FOR EACH COIL AND CORE COMBINATION ***
960 '
980 '   FOR J=1TO35
990 '   L(J)=INT((10^4)*((N(J)/1000)^2)*MHY*U)/(10^4)
1000 '  NEXT J
1020 '
1030 REM **NOW CALCULATE WOUND DIMENSIONS & TEMP RISE USING 30AWG AS AVERAGE**
1040 '
1050 '  WOD=INT(100*SQR((OD^2)-(0.56*(ID^2))))/100
1060 '  WID=INT(100*SQR((ID^2)-(0.56*(ID^2))))/100
1070 '  WHT=INT(100*(HT+ID-WID))/100
1080 '
1090 '  SIDES=3.14156*WOD*WHT
1100 '   TP=(3.14156*WOD^2)/4
1110 '
1120 REM **ASSUMES SMW RADIATED/SQ.IN./DEGREE CENTIGRADE**
1130 '
1140 TEMRIZ=INT(100*(APKMAX(19)^2*RDC(19)*125/(SIDES+TP)))/100
1150 '
1160 REM *** OUTPUTS TO PRINTER ***
1170 '
1180 OPEN 4,4:PRINT#4,CHR$(147);CMD4
1190 PRINT
1200 PRINT
1201 PRINT TAB(25)"***   TORCHK1   PROGRAM    ***"
1202 PRINT
1210 PRINT TAB(25)"*** TORDID INDUCTOR DESIGN ***"
1220 PRINT
1230 PRINT TAB(32)"POWDER IRON CORE"
1240 PRINT
1250 PRINT TAB(36)"CARRYING"
1260 PRINT
1270 PRINT TAB(29)"MAXIMUM  PEAK  CURRENT"
1280 PRINT
1290 PRINT"          ************************************************************";
1291 PRINT"*******"
1300 PRINT
1310 PRINT
1320 X=44:Y=13
1330 PRINT TAB(Y)"CORE O.D.:";SPC(X-LEN(STR$(OD)));STR$(OD)
1340 PRINT TAB(Y)"CORE I.D.:";SPC(X-LEN(STR$(ID)));STR$(ID)
1350 PRINT TAB(Y)"CORE HT.  :";SPC(X-LEN(STR$(HT)));STR$(HT)
1360 PRINT TAB(Y)"CORE PERM.:";SPC(39-LEN(STR$(U)));STR$(U)
1370 PRINT TAB(Y)"MHY/1000T:";SPC(40-LEN(STR$(MHY)));STR$(MHY)
1380 PRINT TAB(Y)"MAG. PATH:";SPC(43-LEN(STR$(MMP)));STR$(MMP)
1390 PRINT TAB(Y)"WINDOW AREA:";SPC(41-LEN(STR$(AW)));STR$(AW)
1400 PRINT TAB(Y)"MEAN TURN :";SPC(42-LEN(STR$(MLT)));STR$(MLT)
1410 PRINT TAB(Y)"HDC PK MAX:";SPC(39-LEN(STR$(MPK)));STR$(MPK)
1420 PRINT TAB(Y)"%MU @ HDC PK MAX:";SPC(36-LEN(STR$(P)));STR$(P)
1430 PRINT
1440 PRINT TAB(Y)"WOUND O.D.:";SPC(43-LEN(STR$(WOD)));STR$(WOD)
1450 PRINT
1460 PRINT TAB(Y)"WOUND I.D.:";SPC(43-LEN(STR$(WID)));STR$(WID)
1470 PRINT
1480 PRINT TAB(Y)"WOUND HT.:";SPC(43-LEN(STR$(WHT)));STR$(WHT)
1490 PRINT
1500 PRINT TAB(32)"TEMP. RISE FOR 30AWG(CU LOSS ONLY)"
1501 PRINT TAB(27)"IS APPROX:";TEMRIZ;" DEG C."
1510 PRINT
1520 PRINT"          ************************************************************";
1521 PRINT"*******"
1530 PRINT#4,CHR$(19);CMD4: REM SKIP TO TOP OF NEXT PAGE
1540 PRINT
1541 PRINT"*****************************************************************"
1543 PRINT TAB(16)"***   TORCHK1   PROGRAM    ***"
1544 PRINT
1545 PRINT TAB(21)"CORE O.D.:";SPC(9-LEN(STR$(OD)));STR$(OD)
1546 PRINT
1550 PRINT TAB(25);U;"PERM."
1570 PRINT "              FULL                                  I"
1580 PRINT "  WIRE       COIL         DC RES       CU       PEAK        INDUC  "
1590 PRINT "  SIZE      TURNS          OHMS        WT       AMPS         MHY  "
1620 PRINT
1640 PRINT"*****************************************************************"
1660 PRINT
1670 FOR J=1 TO 35
1680 PRINT SPC(6-LEN(STR$(IDADE(J))));STR$(IDADE(J));,
1681 PRINT SPC(10-LEN(STR$(N(J))));STR$(N(J));,
1682 PRINT SPC(14-LEN(STR$(ROC(J))));STR$(ROC(J));,
1683 PRINT SPC(8-LEN(STR$(CWT(J))));STR$(CWT(J));,
1684 PRINT SPC(10-LEN(STR$(APKMAX(J))));STR$(APKMAX(J));,
1710 PRINT SPC(12-LEN(STR$(L(J))));STR$(L(J))
1740 NEXT J
1750 PRINT
1760 PRINT
1770 PRINT"*****************************************************************"
1780 PRINT#4:CLOSE4
1790 END
READY.
```

age reduction of permeability desired. The remaining core data is obtained from the selected core data sheet (**fig. 4**).

Enter all inputs and run the program. Outputs are shown in **figs. 5** and **6**.

When zeros are printed, it means that the calculated values are less than the number of decimal places in the printing format. These small values are not practical to measure in most cases.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

meteor scatter communications

It's been two years since my first article on improving meteor scatter communication appeared in this column.[1] That article generated lots of interest and many readers have used the techniques described to improve their own meteor scatter QSO success rate. In that column, I described a method and a computer program designed to help pinpoint the peak of each of the major meteor showers. The predicted peak for upcoming meteor showers you see listed at the end of each monthly column is derived using this method.

Since that column was written, many new improvements in prediction have taken place: shower peaks can now be predicted with greater accuracy, and computer programs that help optimize scheduling are readily available. Because the summer months — along with a short peak from December through early January — are prime time for meteor scatter communications, it seems this would be a good time to present some additional tips on improving meteor scatter communications.

a quick review

Meteor scatter communications uses the ionized trails left by meteors as they enter the earth's ionosphere. There are two basic types of meteors, *sporadic* and *shower*. Sporadic meteors are random in nature. They peak daily at about 6 AM and are at their minimum at about 6 PM local time. Shower meteors, on the other hand, peak at specific times of the year and can begin at any time of the day, depending on the radiant (the place in the sky from which the meteors seem to *emanate*).

Amateurs in North America most often use the shower types, believed to be the remnants of old or extinct comets. These meteors generally provide more ionization and their time of arrival is more predictable than that of sporadic meteors. Shower meteors usually are in greater quantity than sporadic meteors; the typical shower lasts from 12 hours to several days (for at least 25 percent of the peak time). As a result, meteor showers are more reliable for completing a QSO than sporadic types.

"That's fine," you say. "I've looked at the meteor shower peak prediction method in reference 1 and even ran schedules during the peak hours listed at the end of 'VHF/UHF World' each month with only limited success. What am I doing wrong?"

Several questions must be answered. First and foremost, was the other station definitely on the air during the schedule? Was it on the right frequency and using the correct timing sequence? Was all your gear and the other station's gear operational — or did one of the stations have a transmitter or receiver sensitivity problem? Was the transmitted power and antenna gain sufficient at each end of the path? "Of course," you insist. Then read on!

One potential problem is the distance between stations. Generally, the optimum distance for meteor showers is between 700 and 1000 miles (1125-1610 km). Distances less than about 700 miles are a problem because your antenna must be elevated for maximum signal strength and the bursts are usually shorter.

Longer distances out to about 1400 miles (2250 km) are possible using meteor scatter. However, for best results use only the high-speed meteor showers, which have increased ionization higher in the E region. (More on this later.)

The most productive schedules are usually run during the hours or days when the meteors are above 25 percent of the peak of the shower; this is called the *peak time* of the shower. Running a day or so ahead or behind the peak may be less productive, especially if the shower is one of short duration.

The procedure for determining the shower peak time and date was described in detail in reference 1, so it won't be repeated here, though **table 1** has been revised to show the latest ecliptic longitude at 0000 UTC. For those who don't care to work out the peaks or use the computer program listed in reference 1, the predictions of major showers are listed each month at the end of this column.

W9IP mentioned that the ecliptic longitude tables shown in reference 1 may not be accurate enough to be reused every four years as originally suggested. This is true because the tables may vary by up to several hours over a four-year period. So I've updated the daily ecliptic longitude in **table 2** and made appropriate corrections.

If a "fudge factor" of about 0.12 (equivalent to about three hours in four years) is added to the values shown in the tables every four years hence, the tables can be used for many years with a high degree of accuracy. For example, the 1986 table can be reused for 1990 by simply adding 0.12 to each value. Hence January 3, 1990 has a

Table 1.

Shower name	E.L.*	Best dates	Duration**	Hourly rate†	Velocity (km/sec)	Local time‡ rise/set
Quadrantids	282.855	January 2-4	14 hrs	110	41.5	2300-1800
April Lyrids	31.8	April 20-24	2.3 days	15	47	2100-1100
Eta Aquarids	44.5	May 2-7	3 days	20	67	0300-1200
Arietids	76.5	June 1-15	8 days	60	37	0300-1530
June Lyrids	84.5	June 10-21	7 days	10	31	2100-1100
Delta Aquarids	126.0	July 26-30	7 days	35	42	2200-0600
Perseids	139.3	August 10-14	4.6 days	65	60	(Note 1)
Orionids	207.4	October 18-23	2 days	30	67	2230-0930
Taurids	221.0	October 30-November 10	20 days	10	30	1900-0630
Leonids	234.7	November 14-19	4 days	20	71	2300-1230
Geminids	261.2	December 10-15	2.5 days	50	36	1800-1000
Ursids	270.5	December 20-24	2.2 days	15	33	(Note 2)

*Ecliptic longitude in 1950 coordinates.
**Duration is the amount of time that the meteors are above 25 percent of the peak of the shower.
†Estimated meteors per hour at maximum. Can vary greatly from year to year depending on shower.
‡Local time for northern mid-latitudes.

Note 1. Never sets. Minimum at 1730.
Note 2. Never sets. Minimum at 2030.

corrected ecliptic longitude of 282.43; 1994, a corrected value of 282.55. For maximum precision, the computer program shown in reference 1 is recommended, since it contains an algorithm for continuously updating the daily ecliptic longitude with minimal error, regardless of dates.

However, *the primary factor in increasing the probability of a successful QSO is determining the optimum time of the day for the desired path direction during or near* (± 12 hours) the predicted peak of a meteor shower. W4LTU provided guidelines in his tables in references 2 and 3, but they were only guidelines and no specifics on distances or probabilities were included.

Why is this consideration so important? Meteors enter the ionosphere in a specific geometry, depending on where they're encountered by the earth in its orbit around the sun. If the meteors are entering on the opposite side of the earth from your QTH, the chance of a completed meteor scatter QSO is slim.

If the geometry of the meteor shower is favorable, there will be a time when the meteors enter the ionosphere in the optimum direction at the best entry angle, yielding specular (i.e., mirror-like) reflection to radio signals on a specific path as explained by Bain.[2,3]

Villard, et. al., described this geometry in detail in reference 4. Meeks and James further defined the ideal geometry for specular reflection on forward scatter, suggesting that the ideal geometry occurs when the meteors cross at right angles to the great circle path between the transmitter and receiver and enter the ionosphere at an elevation angle of 45 degrees.[5] Meeks and James then derived a complicated set of formulas for determining the decrease in performance from the ideal case. They called this term "effectivity."

This can best be visualized as a mirror reflecting the wave to the proper point. Whenever the trail is properly oriented (consider the reflection region to be the mirror), the incident wave will be reflected directly back to the desired destination. Furthermore, the angle the incident wave makes with a perpendicular to the mirror will be the same as the angle the reflected wave makes with the perpendicular. As the mirror is moved from this ideal orientation, only a portion of the incident wave will be projected onto the correct region of the mirror for reflection back to the desired location.

Thus as the reflection mirror is tipped away from the ideal orientation, the amount of reflected signal in the desired direction falls off. By calculating the amount the reflecting surface is tipped away from the ideal orientation, it's possible to evaluate the "effectiveness" of the reflection process.

Damboldt simplifed the equations for effectivity.[6] With the help of Chip Brown, KR1P, I have simplified these equations even further to yield a percentage of probability from the ideal case as follows:
Effectivity = 200• [Sin H•Cos H•Sin(p-a)]. Where Effectivity is a percentage from 0-100 percent, *H* is the

elevation angle of the meteor entry path, *p* is the azimuth of the great circle path, and *a* is the azimuth of the meteor path.

For example, if the great circle path from the transmitting to the receiving station is 225 degrees, the meteor path is 135 degrees, and the meteor enters the ionosphere at an elevation angle of 45 degrees, the effectivity is 100 percent. However, if the meteor enters on an azimuth path of 180 degrees at a 30-degree elevation angle, the effectivity is only about 62 percent. *Generally speaking, effectivities of greater than 50 percent are desired. If the effectivity drops below 25 percent, the path is very poor and probably not worth your precious time.*

finding the best time for a path

As mentioned above, one of the most important considerations in setting up a schedule is choosing the optimum time for the desired path direction. Some of this information is included in the table prepared by W4LTU.[2,3] However, it is incomplete and doesn't list all the optimum times or all path directions, or cover different distances .

A computer program that calculates the most "effective" time for a meteor scatter path between your station and the station scheduled during any of the well-known annual meteor showers has been written. Even if you don't have a computer, read on; there will be useful information for you!

The program first determines the location (above the great circle path) in the E region where the scattering from the meteor trail will occur, provided that the meteor trail passes through this region in the correct direction. Either of W4LTU's *QST* articles[2,3] illustrate this meteor trail geometry for a given beam heading. The heading to the other station can be easily found using the equations for determining the great circle bearing direction.[7]

The program works out all the astronomical calculations necessary for the meteor shower selected. Armed with these three known quantities (the location of the reflection point, the great circle path, and the direction of the path of the meteors as they enter the E region), the program will evaluate the "effectiveness" (or the orientation of the ionized trails in providing a properly positioned layer or surface on which the specular reflections can occur.)[5]

The orientation of the ionized meteor trails in this particular location in the E region varies as the earth rotates; the program will indicate the optimum or most "effective" time of the day for the use of the path. Because some showers are from radiants of high declination while others are from lower declination, it follows that some showers are better suited for particular great circle paths.

The program also calculates the beam offset angle from the great circle bearing path as a function of time for those who need to compensate for this effect.[2,3]

The astronomical calculations begin with the equations found in the annual *American Epherimis and Nautical Almanac*. These equations are used to determine the day of the year when the earth passes through the shower and the "right ascension" of the shower at the start of that day. This information is then used to determine the direction of the meteors, since they'll be passing through the E region scattering location or volume. Moreover, because of the earth's rotation and its orbit about the sun, this location in the E region will continuously sweep

Table 2. Updated Ecliptic longitudes at 0000 UTC for selected days (from the American Epherimis and Nautical Almanac for the year of interest). The tables can be used over every four years if a small correction is applied as described in the text.

Date	1986	1987	1988	1989
Jan 3	282.31	282.13	281.91	282.68
Jan 4	283.33	283.15	282.93	283.70
Jan 5	284.35	284.17	283.95	284.72
Apr 21	30.61	30.43	31.20	30.97
Apr 22	31.58	31.41	32.18	31.94
Apr 23	32.56	32.38	33.15	32.92
May 4	43.25	43.08	43.84	43.61
May 5	44.22	44.04	44.81	44.58
May 6	45.19	45.01	45.78	45.54
Jun 5	74.06	73.89	74.65	74.41
Jun 6	75.02	74.84	75.60	75.37
Jun 7	75.98	75.80	76.56	76.32
Jun 14	82.67	82.50	83.25	83.02
Jun 15	83.62	83.45	84.20	83.97
Jun 16	84.58	84.41	85.16	84.93
Jul 27	123.69	123.52	124.27	124.04
Jul 28	124.65	124.47	125.23	125.00
Jul 29	125.60	125.43	126.18	125.96
Aug 11	138.05	137.88	138.63	138.40
Aug 12	139.01	138.84	139.59	139.36
Aug 13	139.97	139.80	140.55	140.32
Oct 20	206.31	206.13	206.90	206.67
Oct 21	207.30	207.12	207.90	207.67
Nov 2	219.28	219.09	219.88	219.64
Nov 3	220.28	220.09	220.88	220.67
Nov 4	221.28	221.09	221.88	221.67
Nov 16	233.34	233.15	233.94	233.70
Nov 17	234.32	234.16	234.95	234.71
Nov 18	235.35	235.17	235.96	235.72
Dec 12	259.66	259.48	260.27	260.03
Dec 13	260.68	260.50	261.29	261.05
Dec 14	261.70	261.51	262.31	262.06
Dec 21	268.82	268.64	269.43	269.19
Dec 22	269.84	269.65	270.45	270.21
Dec 23	270.85	270.67	271.47	271.23

throughout the swarm of meteors in an ever-changing direction. The ideal direction of the trails as they pass through the E region will be such that specular reflections will take place over the desired path.

For some paths and some showers, these ideal direction situations may be reached. Though it sometimes lasts for only an hour, this near-ideal situation can persist for many hours in other combinations of paths and showers. For some showers this occurs only once a day; for others, it occurs twice during the same day.

The computer program recalculates all of these geometric changes of the path in the E region with respect to the desired propagation path at every 15-minute interval throughout the day. At each 15-minute interval the "effectiveness" of the path geometry is also evaluated.

The final output of the program is a tabulation and plot of the path's "effectiveness" at each 15-minute interval. Thus the "effectiveness" plot indicates the time of day when astronomical orientation of the meteor trails is most favorable for providing specular reflection along the desired path. The actual value of the plot is a measure of how near specular the reflections will be for the time of the calculation.

computer programs

A copy of this program is being offered for IBM or IBM-compatible PCs at a nominal fee to cover cost of reproduction and mailing.* A floppy disk copy of the program is also available.

Credit for writing this program goes to Chip Brown, KR1P, who spent many hours customizing and debugging it so that the results would be as accurate as possible.

A typical printout of the program is shown in **fig. 1**. This shows a southwesterly path of approximately 900 miles (1450 km) during the Perseids meteor shower from a latitude of 40 degrees north and a longitude of 76 degrees west (the corner of VUCC grid squares FM19, FM29, FN10 and FN20), approximately in the middle of the Eastern Standard Time (EST) zone. This path can be duplicated or correlated later if you obtain the program.

GMT	OFST	Path Effectivity (%)	GMT
0000	1.2	21	0000
0015	1.3	20	0015
0930	14.6	20	0930
0945	15.2	23	0945
1000	15.6	27	1000
1015	15.8	31	1015
1030	15.7	36	1030
1045	15.2	40	1045
1100	14.3	45	1100
1115	13.2	50	1115
1130	11.8	54	1130
1145	10.2	59	1145
1200	8.7	64	1200
1215	7.3	68	1215
1230	6.1	73	1230
1245	5.0	77	1245
1300	4.1	81	1300
1315	3.3	85	1315
1330	2.7	88	1330
1345	2.3	91	1345
1400	1.9	93	1400
1415	1.6	95	1415
1430	1.4	97	1430
1445	1.2	98	1445
1500	1.1	98	1500
1515	1.0	99	1515
1530	1.0	98	1530
1545	0.9	97	1545
1600	0.9	96	1600
1615	0.9	94	1615
1630	0.9	92	1630
1645	0.9	90	1645
1700	0.9	87	1700
1715	0.9	84	1715
1730	0.9	80	1730
1745	0.9	77	1745
1800	0.9	73	1800
1815	0.9	69	1815
1830	0.9	65	1830
1845	0.9	61	1845
1900	0.9	58	1900
1915	0.9	54	1915
1930	0.9	50	1930
1945	0.9	47	1945
2000	0.9	44	2000
2015	0.9	41	2015
2030	0.9	39	2030
2045	0.9	36	2045
2100	0.9	34	2100
2115	0.9	32	2115
2130	0.9	31	2130
2145	0.9	29	2145
2200	1.0	28	2200
2215	1.0	27	2215
2230	1.0	26	2230
2245	1.0	25	2245
2300	1.1	24	2300
2315	1.1	23	2315
2330	1.1	22	2330
2345	1.2	21	2345
2400	1.2	20	2400

fig. 1. This is a printout of the predicted effectivity of the Perseids meteor shower for a southwesterly path from latitude 40 degrees north and longitude 76 degrees west as generated by the computer program mentioned in the text. Time shown is UTC. See text for further use and interpretation.

*A copy of the meteor shower effectivity program written in BASIC for IBM or IBM-compatible PCs is available for $2.00 from Gary Field, WA1GRC, 2 Pluff Avenue, Reading, Massachusetts 01824. A floppy disk with the same program is available for $15.00, which includes cost of the disk and mailing.

Because no real meteor shower enhancement is indicated from 0015-0930 UTC, this wouldn't be a good time to run this path direction. However, note that the path effectivity exceeds 50 percent (the typical minimum criteria as stated earlier) from about 1115 UTC (0615 EST) until after 1930 UTC (1430 EST). A long-duration smooth peak that reaches maximum at about 1515 UTC (1015 EST) is evident.

Figure 1 is also valid by simply scaling the time for schedules where one of the stations is near the 40th parallel but at a different longitude. A station located near a longitude of 91 west can use this path directly by just adding one hour to the times shown. Since the geometry moves westerly at a rate of 15 degrees per hour, a station near a longitude of 106 west can add two hours.

Beam offset is also shown on the printout for those who want to optimize this geometry. W4LTU has explained this principle in references 2 and 3. Normally this isn't important unless you're on the bands above 2 meters or if the gain of your antenna is high (narrow beamwidth). However, every little bit of help is worthwhile, especially on the higher bands.

using the effectivity graphs

Judging from the many letters and questions received, there seems to be a lot of interest in finding the optimum time for a particular path during a meteor shower. The program described will do this for anyone with the appropriate computer. But some Amateurs interested in meteor scatter don't have computers; others are interested only in certain paths. Still others are interested only in some of the typical optimum paths for the major showers.

Tabulating all path possibilities like the one shown in **fig. 1** would be a monumental task. Consequently, I've decided to provide optimum path data for three of the major meteor showers. Since much of the Amateur communications activity during meteor showers in North America has at least one of the stations near the 40th parallel, this is the data I tabulated. Other stations even several hundred miles north or south will often experience similar effectivity. Although it may be shifted an hour one way or the other, there probably won't be a great deal of change if you're running near the peak effectivity points. If you don't have the computer capability, maybe the other station or a friend with the appropriate computer can run a listing for your specific shower and path.

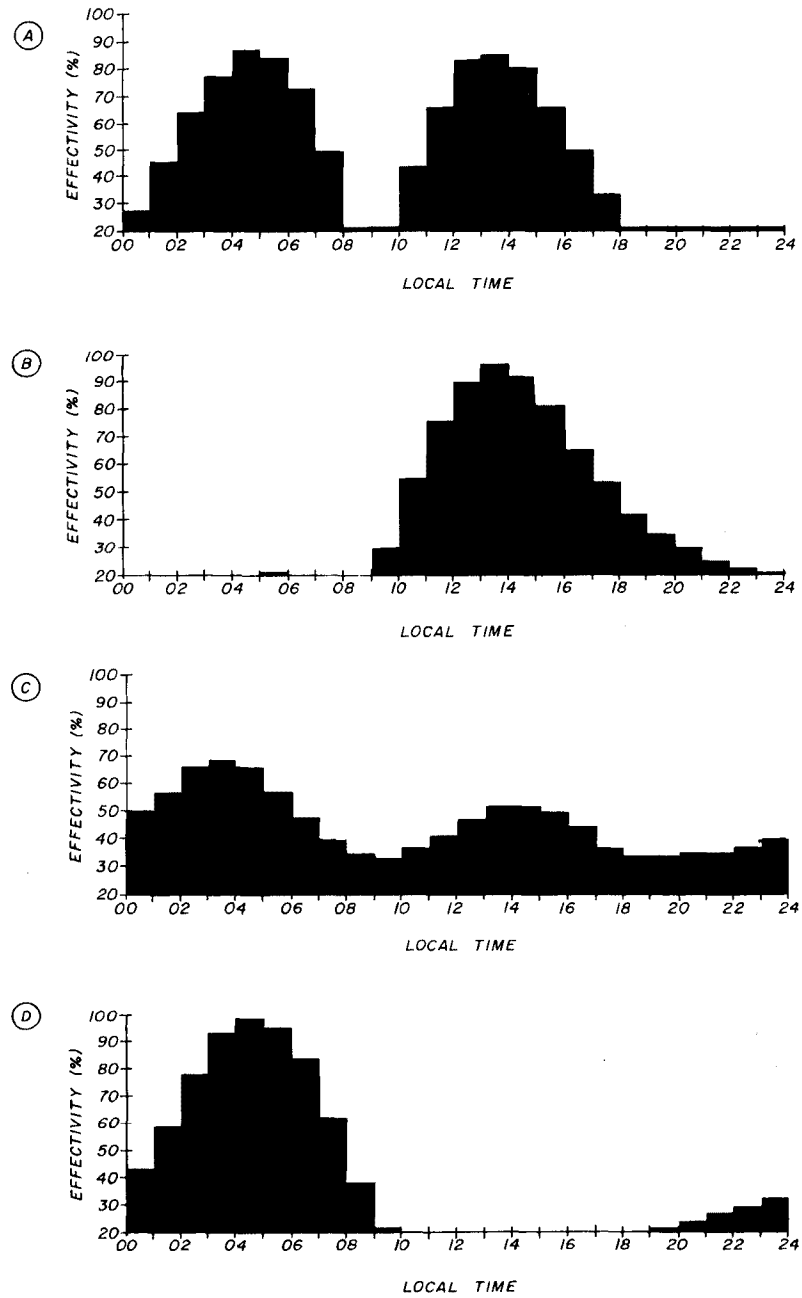


fig. 2. These graphs show the predicted effectivity for the Quadrantids meteor shower. All times shown are local. Original data was taken from latitude 40 degrees north and longitude 76 degrees west but can be used at other locations as explained in the text. Paths shown are as follows: A) north, B) northeast, C) east, D) southeast, E) south, F) southeast, G) west, H) northwest.

The data shown on **figs. 2, 3,** and **4** is for a nominal path length of 900 miles (1450 km). The times listed are local standard time. All the data was run from a latitude of 40 degrees north and a longitude of 76 degrees west, right in the middle of the Eastern time zone of the United States.

Because it's normalized to local time (even though the computer program uses UTC) the data in **figs. 2, 3,** and **4** is directly usable in all time zones. In addition, eight different directions have been shown so that you can pick the optimum time based on the direction desired.

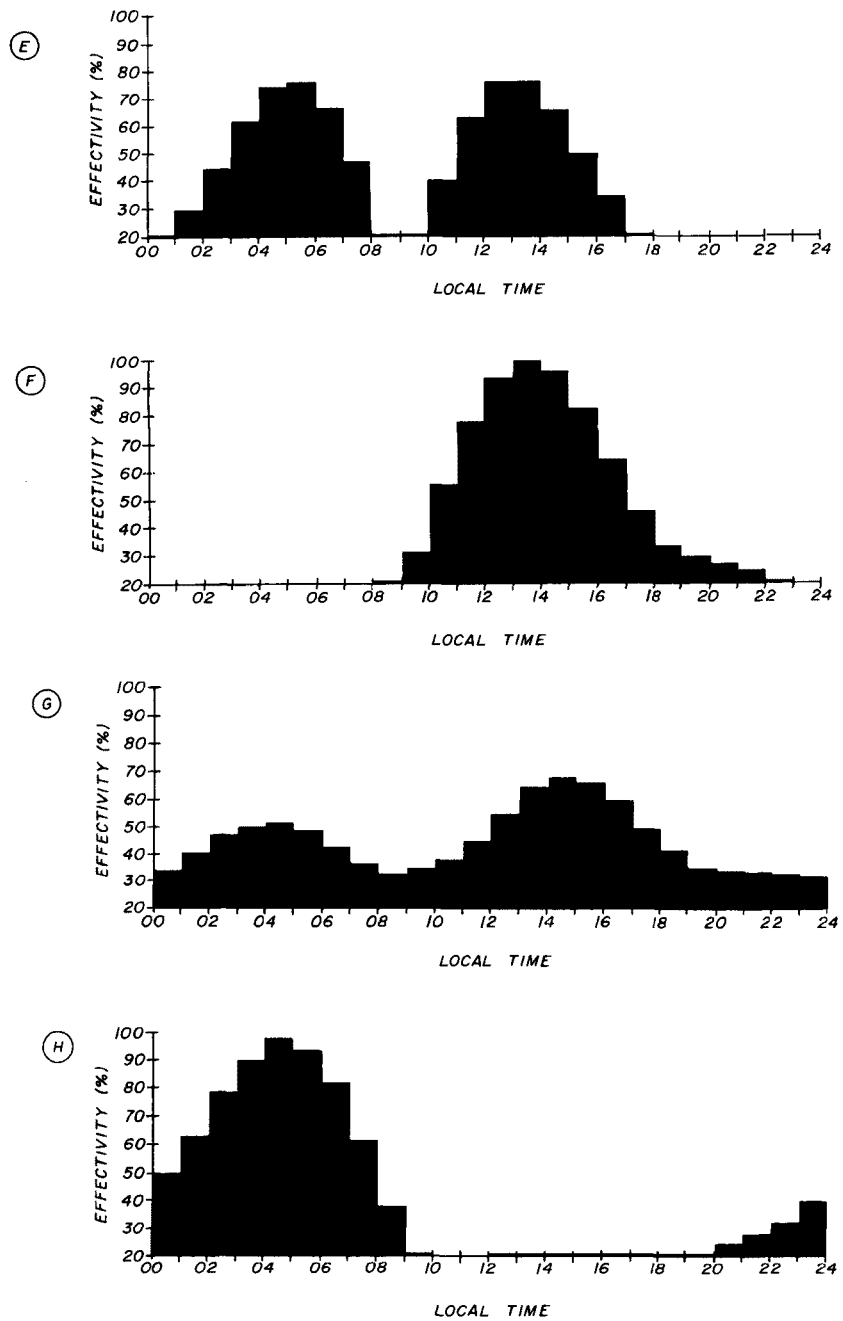


fig. 2. cont.

With these graphs, you'll be able to easily discern any path differences by comparing one with another. For instance, examine the graphs for the Perseids shower (**fig. 3**). You'll see that the east and west paths are fair at two different times per day while the southeast and southwest paths are good only once a day. You'll also notice that the east and west as well as the southeast and southwest paths are identical except for a time shift! However, the latter paths are better when open than the former paths. Other information can be gleaned by examining the graphs for the other paths and showers.

monitoring a shower

Several suggestions about monitoring showers were made in reference 1. Lower frequencies can be monitored. Some Amateurs have suggested 75 MHz because the FAA has thousands of landing beacons on this frequency. Others prefer the FM band, 88-108 MHz.

If you use a lower frequency for monitoring, the lower directivity of your receiving antenna and the presence of bursts don't necessarily indicate that 2 meters or higher frequencies are optimum. Random meteors that will easily ionize up to 108 MHz are often present regardless of the time of day or year. Therefore, I still feel that it's better to use the video carriers on TV channels 12 or 13 as explained in reference 1. These frequencies are good indicators since they also show that the path is open up to at least 200 MHz, simplifying determination of the optimum path.

Furthermore, the speed of a meteor shower and the size of its particles are good indicators of the distance you can communicate. More specifically, because of the earth's velocity of rotation, meteors enter the ionosphere at approximately 15-70 km/second. The entering speed of the particles during a meteor shower is fairly well known and is tabulated in **table 1**.

Meteors typically ionize a fairly narrow region of the ionosphere between 50 and 75 miles above the earth.

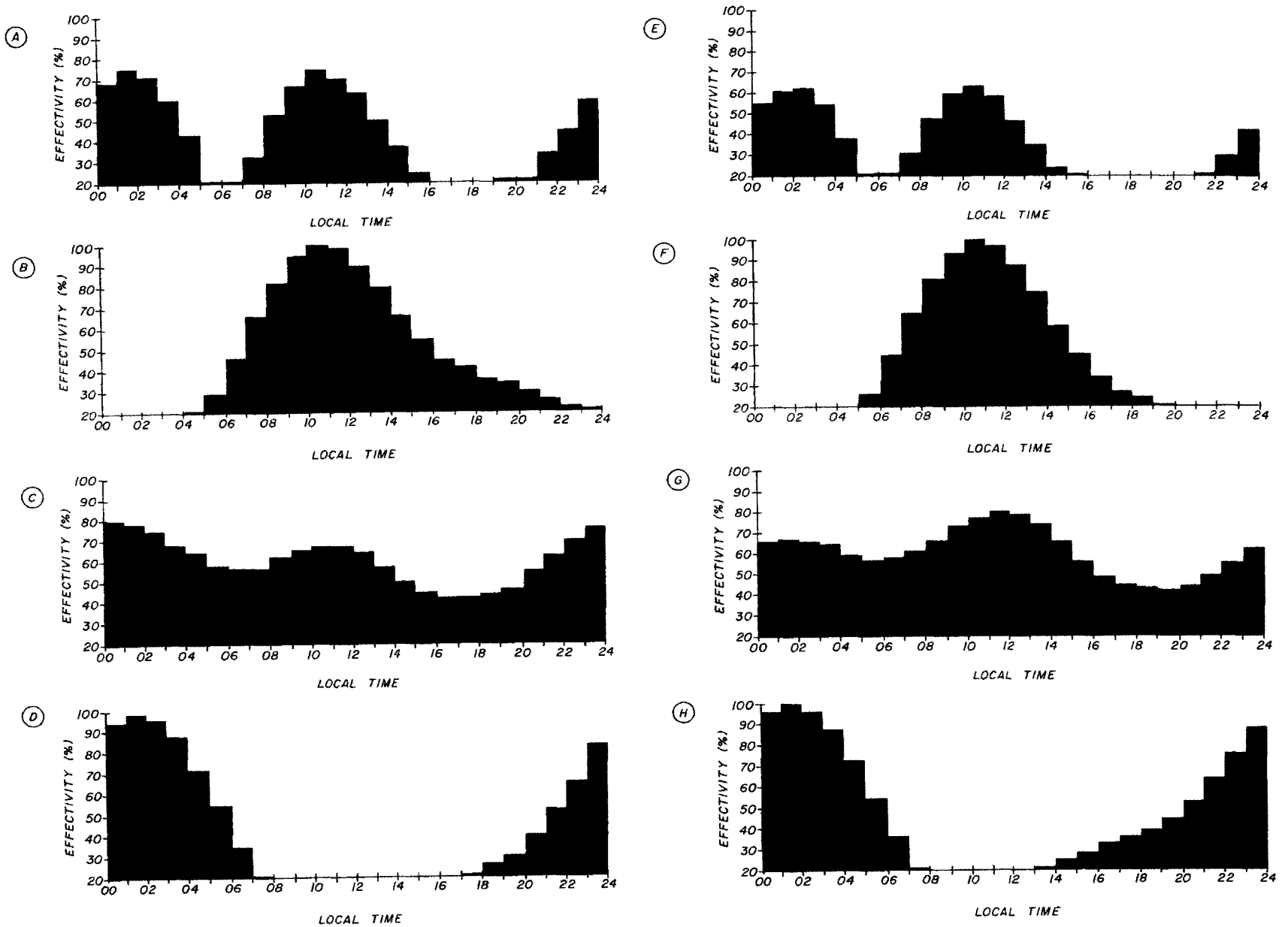


fig. 3. These graphs shown are the predicted effectivity for the Perseids meteor shower. All times shown are local. Original data was taken from latitude 40 degrees north and longitude 76 degrees west but can be used at other locations as explained in the text. Paths shown are as follows: A) north, B) northeast, C) east, D) southeast, E) south, F) southeast, G) west, H) northwest.

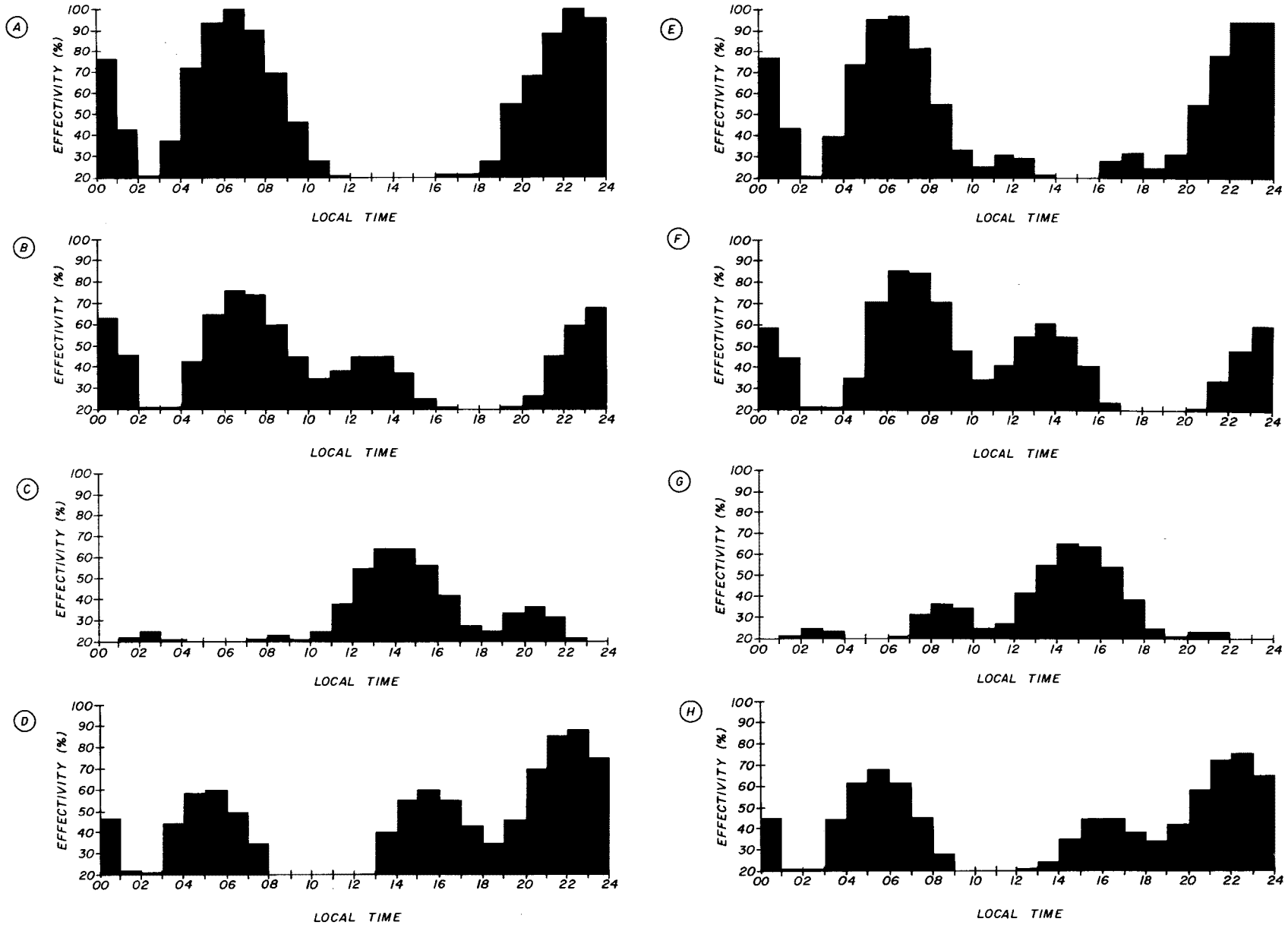


fig. 4. These graphs show the predicted effectivity for the Geminids meteor shower. All times shown are local. Original data was taken from latitude 40 degrees north and longitude 76 degrees west but can be used at other locations as explained in the text. Paths shown are as follows: A) north, B) northeast, C) east, D) southeast, E) south, F) southeast, G) west, H) northwest.

Generally speaking, the higher-speed showers ionize at the upper height and the lower-speed showers at the lower height. Also, those meteors with larger particles tend to ionize in the lower regions. What this means is that the lower the ionization takes place in the ionosphere, the shorter the DX path. The Geminids meteors, for example, have relatively large particles; they're also a relatively slow shower. Consequently, 1100 miles (1770 km) is about the maximum reliable distance.

The Perseids, Orionids, Eta Aquarids, and Leonids, on the other hand, are fast showers and therefore yield the longest communications distances — up to about 1400 miles (2250 km).

I've been trying to integrate this information for some time. Gritting my teeth, I prepared the graph shown in **fig. 5.** While it may not be technically accurate, especially at the upper end, it will at least show a trend based on my meteor shower speed and propagation observations.

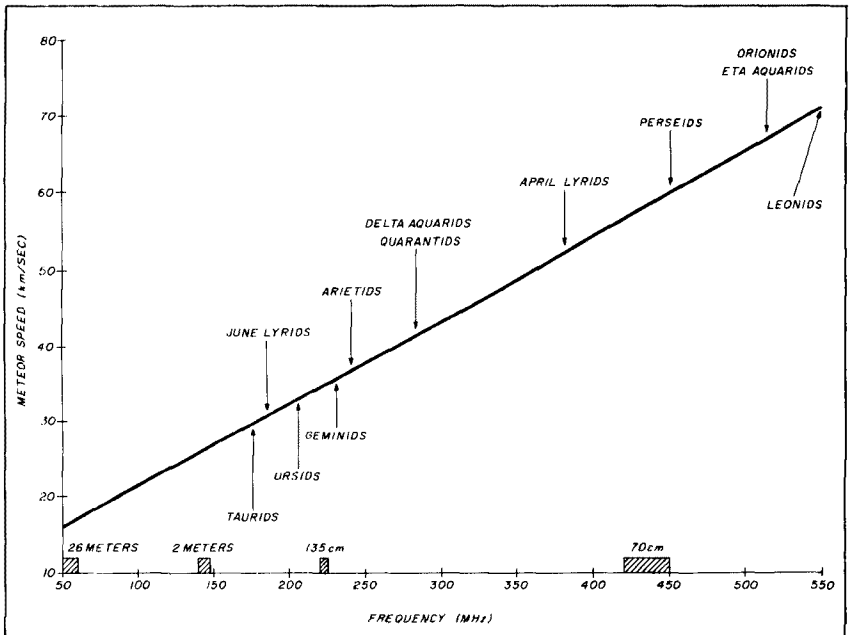


fig. 5. This graphs shows the relative speed of the major meteor showers. By comparing speeds the various showers can be judged to see which will be optimum for your schedules as explained in the text.

From **fig. 5** it should be obvious that the Perseids, Eta Aquarids, Orionids, and Leonids are preferred for 135- and 70-cm operation. However, remember that *lots* of meteors are also desirable. Table 1 shows that among these four showers, only the Perseids has a high hourly rate, a necessary ingredient for completing a QSO. Don't fret. The Leonids are due to peak again in the late 1990s — so be ready!

meteor scatter records

In last July's column, I listed North American meteor scatter records in **table 4.**[8] None exceeded 1470 miles (2365 km). Since then, several stations have sent me reports of being heard further than this distance, but it appears that no QSOs have been successfully completed. All of these contacts were during the Perseids meteor shower.

Ironically the 2-meter meteor scatter record has just been shattered by a great distance and during the Geminids meteor shower! Congratulations to K5UR in Cabot, Arkansas, and KP4EKG near San Juan, Puerto Rico, who on December 13, 1985 completed a meteor scatter QSO during a schedule from 0400-0530 UTC. The length of this path was an incredible 1960 miles (3153 km)!

How can this long distance path exist? Since it took several bursts to complete the QSO and signals were moderate at best, the only plausible explanation is that they used double-hop meteors. This is probable since the quantity of the meteors in the Geminids shower is high and therefore would support such a theory. Does anyone have a better explanation?

summary and conclusion

To improve your chances of successful meteor scatter communications, it's most important to use the optimum time for the desired path based on the geometry of the meteor shower. You can use the graphs and other information generated by the "effectivity"-predicting computer program to determine these optimum times.

Operating during the peak encounter of the shower with the earth is important, but it will be useless if the meteors are on the other side of the earth, if the path of the shower was perturbed by an encounter with one of the other planets in the shower orbit, or if you're not properly oriented to increase the path "effectivity."

Packet radio is now a reality; 1986 should be the year in which many 2-meter QSOs will be completed using this relatively new mode. Europeans have been using a type of packet radio for many years in the form of high speed CW, sometimes exceeding 100 words per minute with great success — even in the absence of showers and at almost any hour of the day. At this high speed the bursts may be short, but if they're tape-recorded and played back at lower speed, a lot of information can be extracted from even a very short burst.

acknowledgments

This month's column would not have been possible without the help of Chip Brown, KR1P, who provided the meteor scatter effectivity program discussed. I'd also like to thank Gary

Field, WA1GRC, who has volunteered to distribute this program, and Mike Owen, W9IP, who pointed me towards new references that yielded improved meteor shower data.

important VHF/UHF events:

June 6: *Predicted peak of the daytime Arietids meteor shower at 0100 UTC*

June 7-8: *ARRL VHF QSO Party*

June 15: *Predicted peak of the June Lyrids meteor shower at 2100 UTC*

June 21: *EME perigee*

June 21: *Mean date ± one month for the peak of sporadic-E propagation*

July 2: *± 2 weeks — look for European openings on 6 meters.*

July 18-20: *CQ Magazine VHF WPX contest*

July 19: *EME perigee*

July 20: *Look for 2-meter sporadic E ± 2 weeks of this date.*

July 25-27: *Central States VHF Conference in St. Louis, Missouri. (Contact NOIS for details.)*

July 29: *Predicted peak of the Delta Aquarids shower at 0845 UTC.*

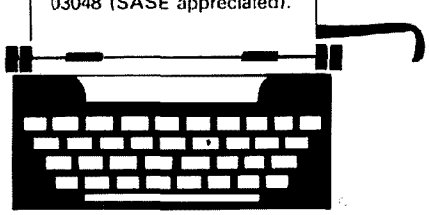
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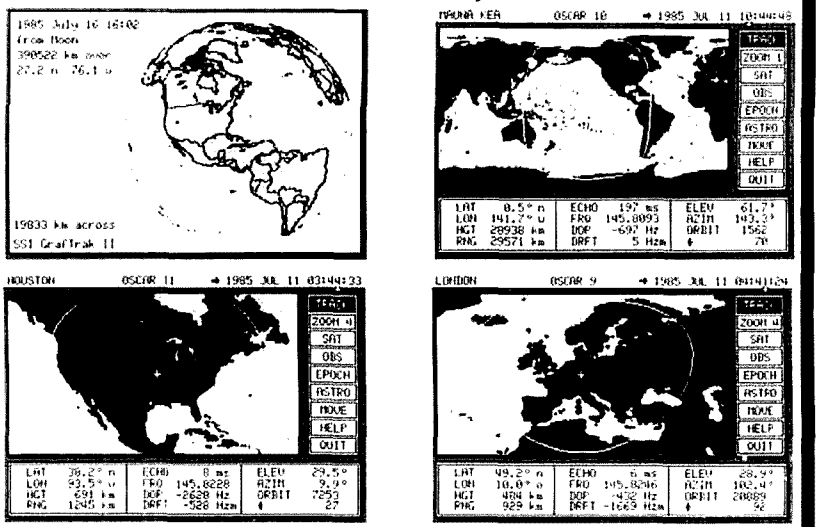
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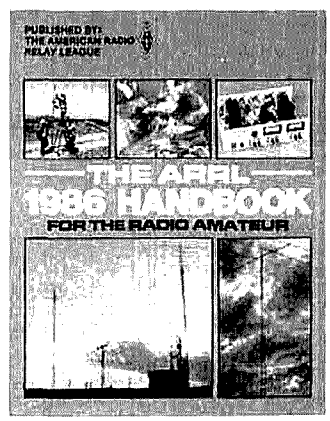
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Experiment with RF circuits using this fast, simple technique

a universal analog breadboard

Even at its best, breadboarding is always a problem, particularly when a little circuit complexity is involved. Digital breadboarding has been made easier by the availability of various commercial breadboards that nicely accommodate the digital DIP packages; analog breadboarding, however, remains an art of arranging, in seemingly random fashion, a multitude of varied components on some substrate material.

Although digital breadboards can be used for analog breadboarding, the component leads just don't seem to fit properly and the interelectrode capacitance is much too high for analog circuits operating much above 1 MHz or so. Other materials such as perf-board work well for anyone who's organized enough to fully lay out the assembly carefully before starting, but if you're going to go to all that trouble, you might as well lay out a PC board. Also, if the circuit being breadboarded is RF in nature, a ground-plane substrate is necessary; this requires tedious cutting and peeling off the ground plane to fabricate a functional breadboard.

The primary value of the breadboard lies in its ease of preparation and use; if a great amount of planning is required for just building the breadboard, much of the advantage of its use is lost.

An analog breadboard system has to be versatile enough to accommodate the wide variety of analog component types, yet provide for very simple fabrication of the circuit. Stray capacitance and inductance must be made very low; good ground plane construction is a must. Changes, additions, and deletions must be simple to make.

choosing a breadboard type

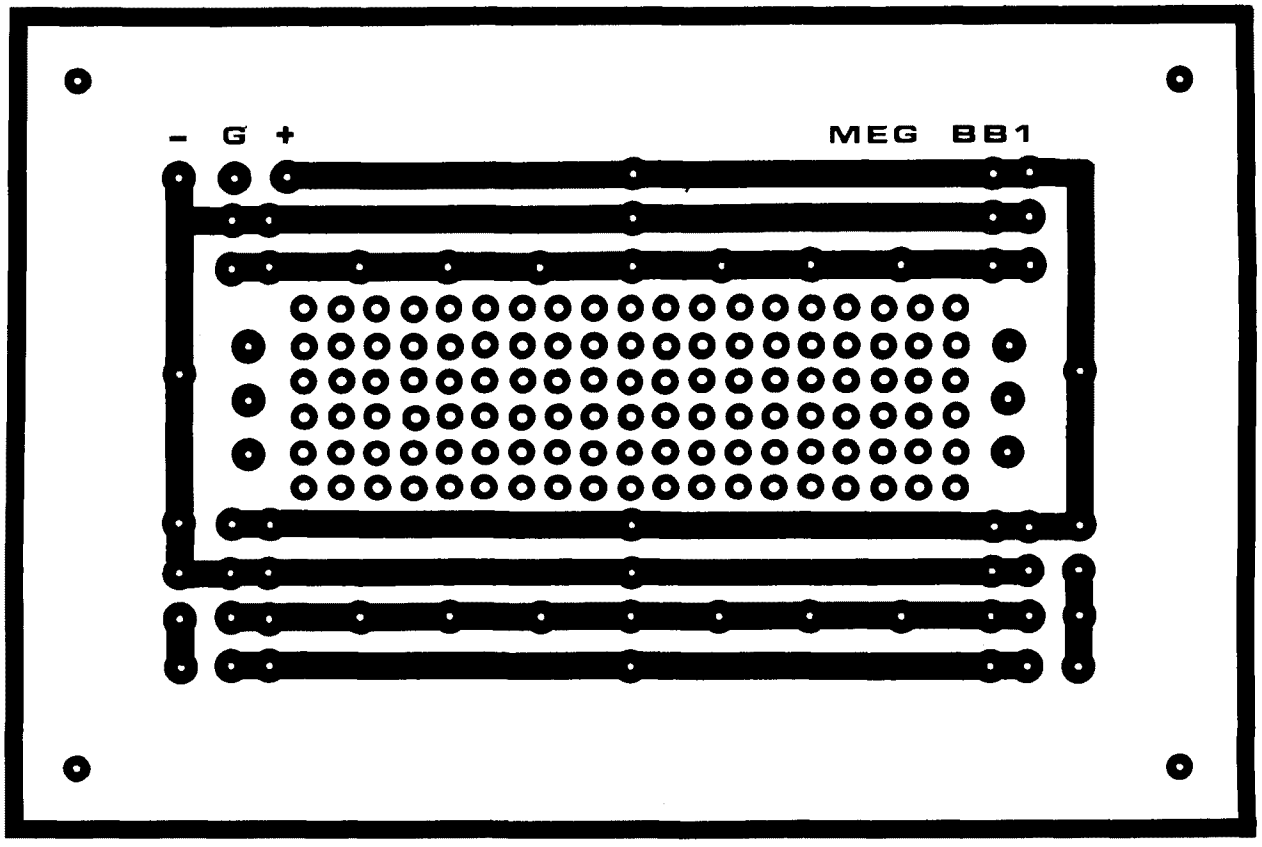
Having tried a variety of breadboards — from commercial spring-boards to nails driven into a piece of plywood — I thought a slightly different approach might be in order. One of my principal requirements was that the breadboard be adequate for simple RF designs. Generally, complex RF circuits such as complete receivers or high-gain amplifiers require more elaborate construction than a simple breadboard will allow because of the difficulty in controlling unwanted coupling. However, crystal oscillators, single-stage RF amplifiers, and modulators, for example, should be appropriate applications for this breadboard technique.

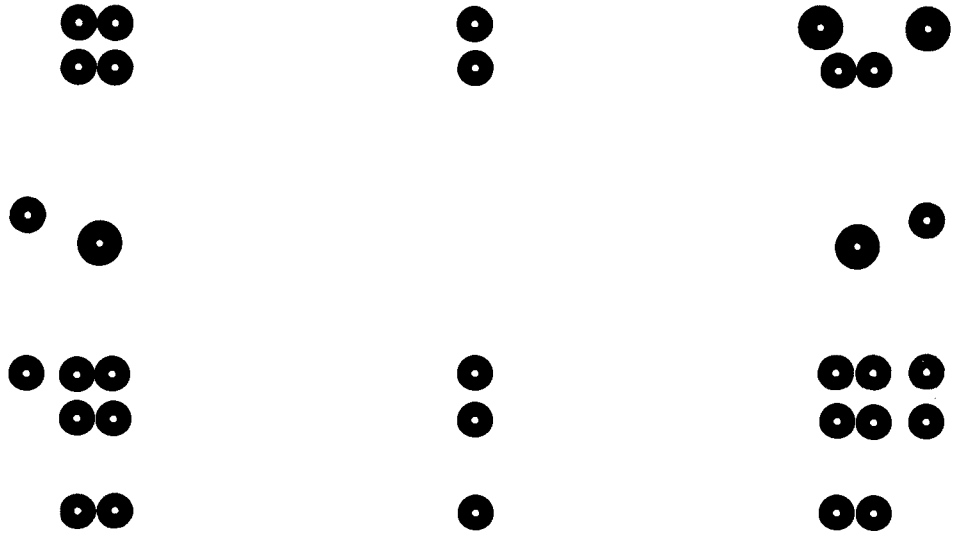
The artwork for a good basic design, a 1/16 inch (0.16 cm) thick PC board with a ground plane on the bottom side and dot and trace pattern on the top side, is shown in **fig. 1**. The traces allow two power supplies and ground to be conveniently bussed around the breadboard. Running power to all the various points where it's needed always seems to present a problem in any breadboard, particularly when you're making a lot of changes as you try out different circuit configurations. The power busses provide a convenient means of accessing the power supplies without having to daisy-chain the power conductors.

Grounding in analog designs is particularly important. Daisy-chaining ground can lead to some very unusual performance (for example, when your amplifier works better as an oscillator than an amplifier). The ground bus provides a good ground for all but the most demanding circuit designs. Power and ground are brought onto the breadboard at the three terminals at the edge of the board. This keeps the power leads out of the circuit area, preventing disturbances that could be caused by the leads lying in the circuitry. It also helps prevent power leads from deciding to detach and fall on the most critical (or expensive) components, generally with spectacular results.

A set of pads for installing power supply bypass capacitors has been provided at the lower left and right-hand corners of the trace pattern. Since long leads are often used to supply power to an experimental circuit, good bypassing of the supplies on the breadboard is very important. The bypass capacitors are often rather large and inconvenient to have in the middle of the breadboard circuit. Placing them at the lower edges of the work space gets them out of the way.

By Michael Gruchalla, 4816 Palo Duro N.E., Albuquerque, New Mexico 87110





GND

fig. 1. Breadboard artwork.

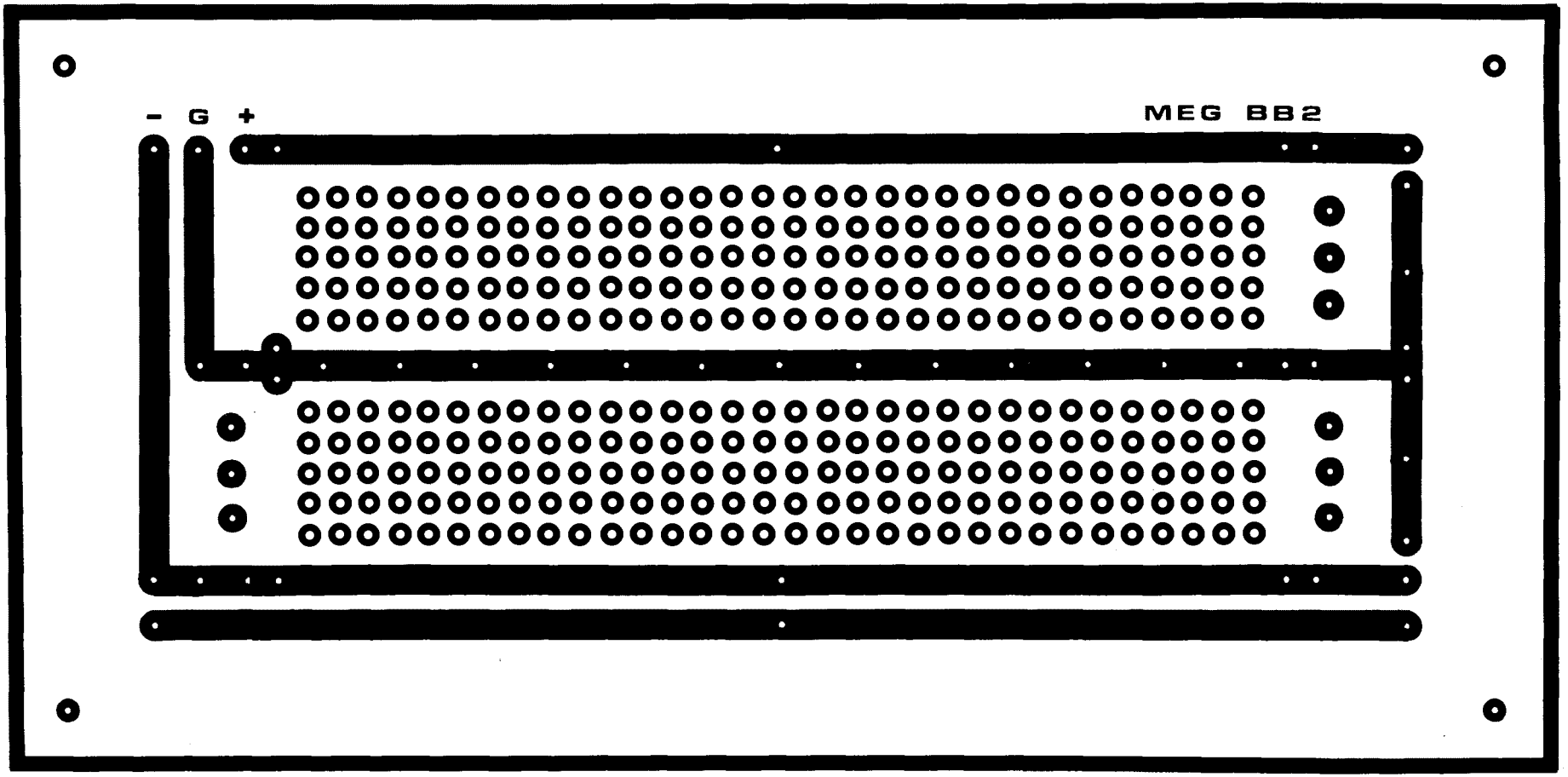


fig. 2A. Artwork for trace side for the larger breadboard.

The actual circuit nodes are simply PC dots arranged in rows and columns, allowing circuit components to be soldered directly to them without the need for holes to accommodate the component leads. The particular dot used has a very large center hole compared to normal PC dots of its size; this minimizes the capacitance of the dot to the ground plane and provides a dot large enough to solder to. The capacitance of each of the pads to the ground plane is about 0.5 pF, which is generally low enough to be usable up to 1 GHz.

Circuit ground can be provided in two ways. The ground bus is adequate for DC and low-frequency applications. High-frequency circuits generally require very tight grounding of various points, and in these cases the ground bus cannot be used very effectively because of the inductance of the wire that must be tied from the node to be grounded to the ground bus. Good grounding can be obtained by drilling through the board in the center of the pad that must be grounded, running a piece of bus wire through the hole, and then soldering it to both the ground plane and the pad. This grounding technique will provide good performance in circuits operating at frequencies as high as L-band (1.0 GHz).

breadboard construction

A few pointers will help you build a more professional-looking breadboard. **Figure 1** shows 1:1 art work for one version of the breadboard, a small unit that's quite convenient for single-stage amplifiers, oscillators, and other small projects. **Figure 2** shows artwork for a larger version with the same basic design but with different features. Art for both the trace and ground plane sides are shown. If you don't have the capability for making double-sided PC boards, you can omit the ground plane art, leaving a full ground plane on the bottom side. You can then use a drill bit (about 1/4 inch) to cut away the ground from the few penetrations where needed — at the power terminals, for example. Use the bit by hand and cut just deep enough to remove the cladding. Though doing this requires some care, it's really quite simple.

The breadboard should be fabricated from high-quality PC board material. Epoxy-glass materials with 2-ounce copper are about the best. The pads will lift off very easily if an inexpensive material such as a phenolic base is used; eventually, after much use, the pads will tend to lift off even the best materials, but if you start with quality material, the pads shouldn't begin to separate until it's about time to retire the breadboard anyway. If possible, the board should be tin-lead plated to minimize oxidation of the conductors with time. After all, old breadboards may lie around in the junk box for years before being resurrected for a quick weekend project, and they're certainly expected to be as useful at that time as they were when they were new.

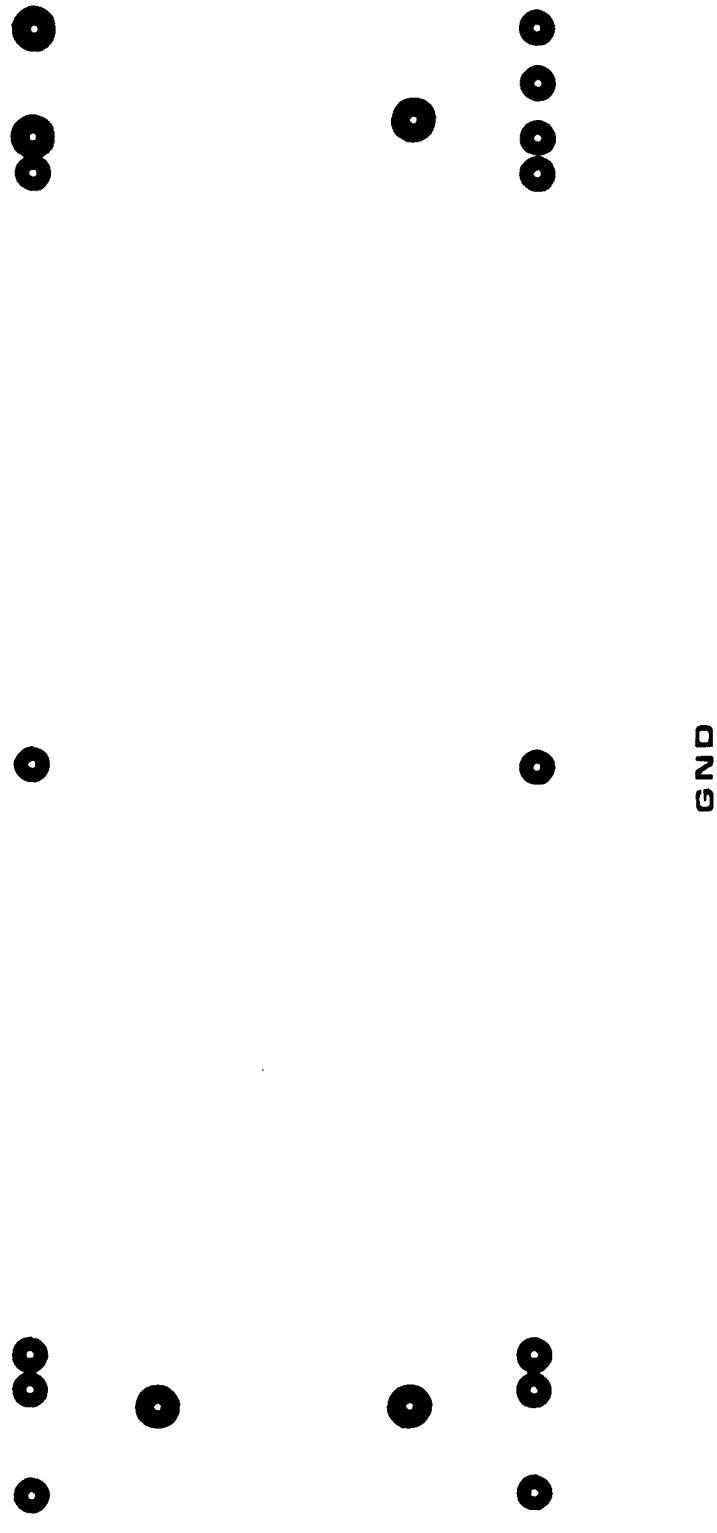


fig. 2B. Artwork for the ground side of the breadboard.

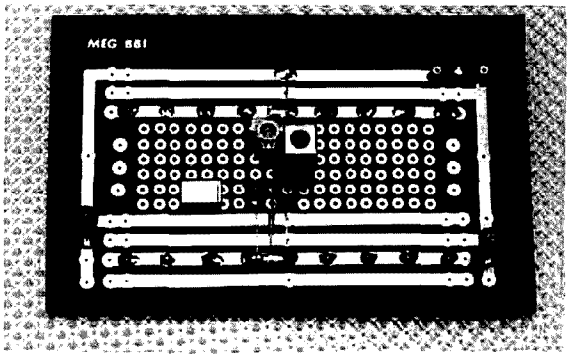


fig. 3. Crystal oscillator example.

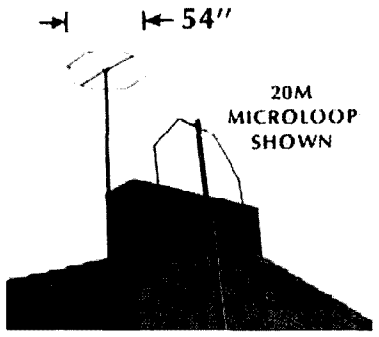
If you have the breadboard fabricated commercially, don't have any holes drilled; this will reduce the cost and provide a good general-purpose board.

Before using the breadboard, tend to a few details to assure professional results. The three power terminals should be mounted at the indicated pads (+, G, and-). Drill a suitable hole through the center of the pads and swage in the terminals. Solder the two power terminals at the pad and the ground terminal at both the pad and ground. If you can't find terminals, you can use a loop of heavy bus wire; about 18 gauge works well. Solder the bus wire to the ground pad and ground plane. Make about a 1/4-inch (0.64 cm) loop in the wire and solder the end back to the ground pad. At the two power pads, solder the bus wire to the pad and loop it around to the corresponding conductor, making about a 1/4-inch loop, and then solder it to the trace. This will help prevent the power pads from pulling off because of the stress exerted by the external power leads. Next, drill holes in the bypass capacitor mounting pads and mount the two bypass capacitors. (A good choice of value is a 47μF, 35 V tantalum.) Be sure to observe the capacitor polarity when mounting. Even if you don't intend to use both supplies, at least drill the holes — you may eventually need an additional power supply, bias supply, or something similar, and having the holes there will prevent shorts that could be caused by metal chips trapped in a completed circuit. The two ground conductors aren't committed to be used as ground. They can be used for any type of bus, but at least one should be used as a ground. The lower is generally the most practical one. Drill holes in all the pads in the chosen ground bus, place bus wire in each hole, and solder to both the ground plane and the pad. Solder the ground plane first. (If you were to solder the pad side first, the piece of wire would probably fall out, since more heat is needed at the ground plane.)

Now, since several solder connections have been made to the ground plane, the board won't sit nicely on your work bench and the wire pieces are likely to scratch your work surface — not a desirable result if your work surface happens to be the dining room

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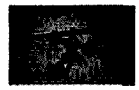

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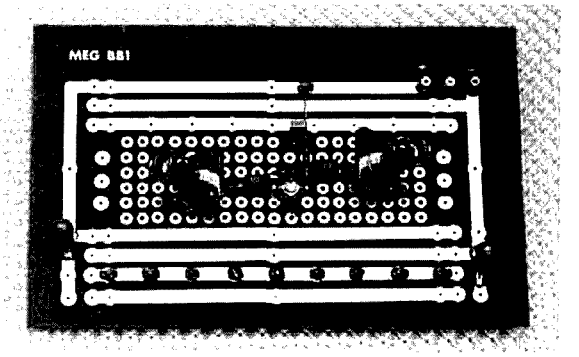


fig. 4. RF amplifier example

table). Adding four small rubber feet to the corners of the board eliminates this problem and gives the breadboard a pleasing appearance and better stability. Either self-adhesive feet or standard screw-mounted feet can be used. Small pads are included at the corners of the breadboard as markers for drilling foot mounting holes. Even standoff spacers can be used with rubber feet to provide space under the board if needed. For example, you could mount the bypass capacitor and power entry terminals under the board to provide more working space on the top side. You could even stack several breadboards together with spacers to form a multicircuit system.

Several extra pads and a bus are included on the artwork. Suffficient space is allowed around the edges of the board to allow for mounting of potentiometers, variable capacitors, connectors, and similar items.

There's nothing magic about the breadboard design, but I find this one quite useful for many analog circuits. **Figure 3**, for example, shows a crystal oscillator operating on TV channel 4 in a television interface application. **Figure 4** shows an amplifier that has a bandwidth of 100 Hz to 1.2 GHz and 15 dB gain; this amplifier was assembled to demonstrate that the breadboard can be used to L-band. From these figures, it can be seen that although these are only breadboards, they make an attractive finished design. Carefully completed and nicely packaged, a relatively handsome finished product can result.

Don't be afraid to experiment with the basic design to find the best configuration for you. The variety is limitless and the configurations shown here are just two possibilities. Remember, for analog breadboards the most important features are low stray capacitance and inductance and good grounding and bypassing of the supplies at the board. Other features such as pad arrangement and size, number of nodes, bus structure, and space for peripheral elements are very much a matter of the specific requirements of the circuit and your personal preference. The basic design presented here should prove to be a good basis for other configurations as well.

ham radio

PRACTICALLY SPEAKING ... Joe Carr K4IPV

selecting and using the right multimeter

Sooner or later almost everyone with more than a minimal interest in electronics begins to look at the various multimeters now on the market. These instruments are more than an aid in troubleshooting — they can be your right arm! A multimeter is a combination instrument that offers the benefits of several meters in one: a DC voltmeter, DC milliammeter, AC voltmeter, and an ohmmeter (plus other functions on certain premium-priced instruments). This month we're going to talk about the various instruments on the market, and what you should look for when selecting a particular one. Next month we'll examine some typical applications and problems.

types of multimeters

Over the years several different forms of meters have been developed. There's also more than one way to classify meters. For example, we could classify them by their active elements: none, vacuum tubes, FETs, and so forth.

We could also classify them according to the display mechanism: analog or digital? Analog meters use a regular pointer-type meter movement based on either the D'Arsonval or Taut-Band designs (there are others, but they're found only rarely, if ever, in multimeters). Their scales are printed on the panel behind the movable pointer. The digital meter, on the other hand, uses digital techniques to make the measurement and displays the result on digital numerical readouts (either LED or LCD).

world's oldest multimeter

Figure 1 shows a modern example of perhaps the oldest form of multimeter: the Volt-Ohm-Milliammeter, or "VOM." Based on an analog meter movement, this meter has switch-selectable ranges for DC volts, AC volts, DC milliamperes and resistance (ohms). Classical VOMs don't contain any active devices and don't require any power at all except for a single battery for measuring resistance.

Sensitivity is a determining factor in the purchase of such an instrument. The "sensitivity figure," a measure of the voltmeter's input impedance, tells us how much load the meter places on the circuit being tested. In the case of **fig. 1**, the sensitivity is 20,000 ohms/volt. If we multiply the full-scale voltage on any given range by the sensitivity, we find the input impedance. For example, when the 50-volt scale is selected, the input impedance is (50 V x 20,000 ohms/V), or 1,000,000 ohms.

The disadvantage of the VOM is that it sometimes loads circuits too much for practical use. In vacuum tube circuits, the meter works fine most of the time; in solid-state circuits, however, there are often problems. For example, when measuring the base voltage of an NPN common-emitter transistor amplifier (if operated from a single positive power supply), we expect to find a base-to-emitter voltage of 0.2 to 0.3 for germanium and 0.6 to 0.7 for silicon transistors. So, using the 1.5 volt DC scale, we find that the meter input impedance is (1.5 V x 20,000 ohms/V), or 30,000 ohms. Unfortunately, this resistance is on the same order of magnitude as bias resistances in transistor circuits, so the application

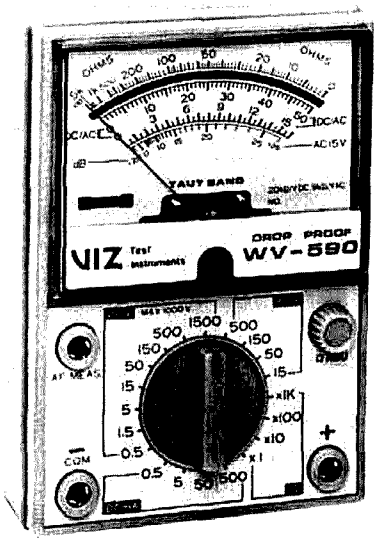


fig. 1. The oldest form of multimeters — the VOM.

of the meter probes changes the bias conditions — and renders the measurement invalid.

The advantage of the VOM is that it is truly a minimum-hassle instrument to keep, store, and use. It's also very portable. These factors make the VOM popular with people who work for long periods in remote areas without easy access to fresh batteries.

Finally, people who work around high-power RF amplifiers might want to keep one of these instruments in their tool kits even if they normally use a modern digital instrument. The classical VOM, without any active devices to be misbiased by strong RF fields, will work in the presence of transmitters, where other, more expensive instruments might fail.

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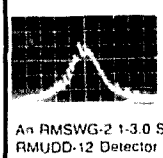
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FET multimeters

Figure 2 shows another form of multimeter, the field effect transistor (FET) mulitimeter (FETVM or FETMM). This type is representative of a larger class of electronic voltmeters, and is a member of the class that includes vacuum tube voltmeters (VTVMs) and transistor voltmeters (TVMs). These devices are sometimes referred to collectively as electronic voltmeters, or EVMs. The names for these instruments are derived from the input circuit device: vacuum tube, transistor, or field effect transistor.

The purpose of the input amplifier device is to increase the input impedance, thereby reducing circuit loading. For example, a typical specification for the VTVM was an input-Z of 10 megohms, with an additional 1 megohm in the probe, for a total of 11 megohms. There is usually also 50 pF of capacitance shunting the input resistance. Modern FETVM and other EVM types sometimes sport input impedances even higher than 11 megohms – I can recall seeing 100 megohms advertised.

Like the VOM, the FETVM is a multirange meter that uses a front panel switch to select range, although in this case the functions (DC volts, AC volts, ohms, etc.) are selected by pushbutton switches. Unlike the VOM, the FETVM (and all EVMs) require DC power supplies for their active devices. In most cases, the DC power supply is a battery, although some 115 VAC line operated models are available. Unlike the classical VOM, EVM instruments are somewhat sensitive to RF fields, so many of them will not work well around high power transmitters and other RF generators such as electrosurgery machines, diathermy, or inductive heating devices, for example.

Both the FETVM and VOM have an AC volts scale. Even where there's a low range scale for the AC mode, the AC voltmeter in these instruments is not a substitute for the AC voltmeter called for in the measurement, alignment, and troubleshooting of audio circuits and devices. The AC scales in these multimeter instruments are generally accurate only at power line frequencies. Beyond an upper limit of 400 Hz or so, the accuracy of these instruments falls off rapidly as frequency increases.

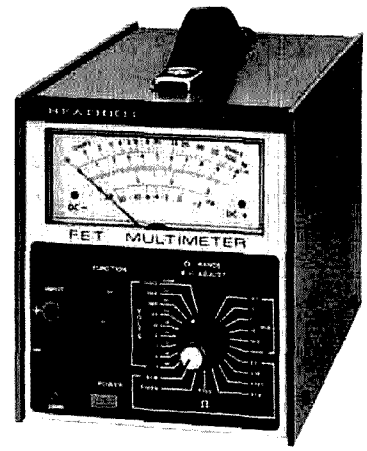


fig. 2. High input impedance is – provided by the FET multimeter.

Examples of some representative digital multimeters are shown in **figs. 3 and 4**. These instruments use an internal analog-to-digital converter to convert the input voltage to seven-segment digital display output. Like the EVM and VOM, the digital multimeter (DMM) offers DC volts, AC volts, milliamperes, and ohms ranges.

what's a "half-digit?"

You'll see DMMs advertised as "2-1/2 digit," "3-1/2 digit," or "4-1/2 digit." Care to guess what a "half-digit" is? Because the most significant digit (the one all the way to the left) can be just 0 or 1, it's billed as a half-digit. Consider, for example, the 3-1/2 digit instruments. The basic range of these instruments is 0 to 999 units, with 100 percent overrange, for a total of 0 to 1999 units. The number of digits is a rough measure of precision, although a load of digits doesn't necessarily guarantee a quality meter. Generally, most Amateur and hobby applications require no more than 3-1/2 digits. There are cases in which 2-1/2 digits isn't enough, yet the 4-1/2

digit instrument may be more expensive than warranted by the application. Emotionally, I prefer to have as many digits as possible, but when it comes to laying out my bucks I prefer to pay due regard to the realities of measurement making — and economics. Dollar for dollar, quality for quality, there is an optimum instrument for each application.

But don't more digits mean more accuracy? And isn't it true that DMMs are in fact inherently more accurate than VOM and other EVM instruments? No! In both caes, the statement is false. The accuracy of the meter depends less on the number of digits in the display than on the quality of the internal workings of the instrument. For example, the accuracy of the internal A/D converter and its voltage reference have considerable effect on accuracy. In general, ethical manufacturers don't display more digits than they can support, but the size of the display cannot be taken at face value as an indicator of meter quality.

Now, regarding the really big heresy: "Whaddya mean my Digital Frammitzenfritzer DMM isn't necesarily more accurate than your crummy old VOM?" Please, before you boil me in oil, let me assure you that I *am* digitally oriented. I sit here writing this article on an Apple IIe personal computer while my IBM-PC sits on the other side of the room, chattering out Chapter 8 of my next TAB book. While it's true that DMMs have the potential for greater initial and long-term accuracy than VOMs, it depends on the implementation, the quality of internal circuits and other factors. There's also the fact that noisy signals tend to be heavily integrated (i.e. low-pass filtered) by the damping inertia of the analog meter pointer, but these same signals will fool many digital meters.

In regard to accuracy and resolution, I recommend buying from a reputable manufacturer and *reading the spec sheet BEFORE you plunk down the money*. Also, be wary of over-spec'ing the instrument. When you're in the market for a troubleshooting instru-

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ment, don't let that gleam in your eye as you view the high-priced super-accurate laboratory models reach down as far as your wallet. Instruments such as those shown in **fig. 4** are very much in evidence in any electronic parts store you visit. Various low-priced models that not only make the classical measurements of the VOM, but do certain other things as well, are available. For example, I recently bought a handheld DMM that measures capacitance. The capacitance measurement scales run from 2000 pF full-scale (with 1 pF resolution) to 2000 μF full-scale. In addition, that model is ruggedized enough to withstand my clumsy work habits . . . no small attribute for *any* practical instrument!

Other features to look for on DMMs are the "Diode" or "High Power" mode and a continuity tester with an aural indicator. The Diode mode is used to test PN junction diodes and bipolar transistors (NPN and PNP). The normal ohmmeter mode in DMMs uses a very low voltage, too low to forward-bias PN junctions. The Diode mode (usually indicated by a diode symbol on the function switch) is a resistance mode, but with a source voltage that's high enough to forward-bias PN junctions. The DMM can be used to measure resistances in-circuit while in the low-power mode, and then test the semiconductors out of circuit in the Diode mode.

The aural continuity tester is basically a resistance scale that goes "beep" when the meter sees a low resistance across the probes. Why is this feature important? Try ringing out a multi-conductor cable while managing the probes in your fingers. Or worse yet, try ringing out an intercom cable while standing on a ladder — or a transmission line while you're hanging off a tower. However clever you may be, you simply can't watch the meter scale and manage the probes — or your own safety — at the same time. The "beep" of the meter tells you when the connection is made, and its absence warns you of an open circuit.

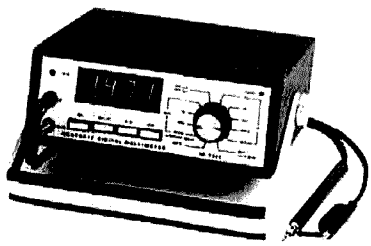


fig. 3. Bench-type digital multimeter.

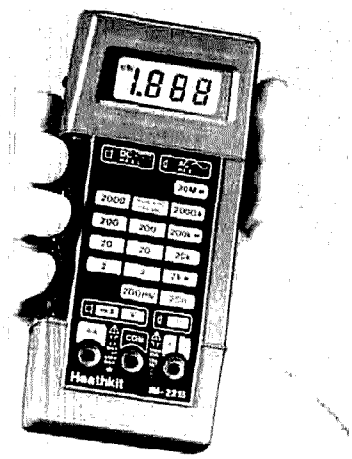


fig. 4. Handheld digital multimeter.

which one to buy?

"OK, Carr, so you're a Certified Bright Guy (CBG) writing for a hot-shot national ham radio magazine — what kind of multimeter did *you* buy?" Believe me, I wish it were that easy! What *you* should buy depends on what kind of work you do, and to some extent, on what you can afford. Fortunately, the cost of a good quality troubleshooting-grade DMM instrument is way down from what it was a couple of years ago. Almost everyone can afford to buy one of those instruments.

Let me describe what I own and what type of work I use them for. The work consists of troubleshooting electronic equipment occasionally and developing small projects for Amateur Radio and electronic hobbyist magazine articles.

Because I have a ham radio set with a 2 kW linear amplifier, it's possible that the meter will have to work in high RF environments. As a result, I own a classical VOM that's almost older than I am. The old VOM has a 25 kV probe as an accessory, although I loaned it out and can't remember who borrowed it. I also own a handheld Beckman DM-25 that measures capacitance in addition to the other "normal" parameters. This last feature turns out to be of immeasurable value in developing projects for publication.

If I were doing any extensive audio work, then I'd probably also buy an AC voltmeter of audio grade. I suspect that most electronic workbench-oriented readers will opt for a small collection of meters that nearly matches my own . . . . **ham radio**

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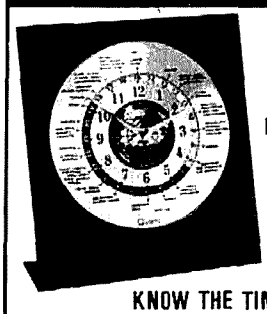
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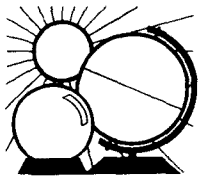
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DX FORECASTER

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sporadic E DX

In last month's column we discussed the characteristics (in terms of location and propagation) of Sporadic E, or E_S. This month, let's take a look at how Sporadic E (E_S) propagation can be used for DX operation.

Because E_S propagation normally provides short-skip conditions — i.e., 1000 miles (1600 km) maximum hop length, with typically one or two hops at best, the range of communications is limited in this mode. The take-off angle must be low (5 to 10 degrees) in order to obtain the maximum distance per hop. Signal attenuation increases with more hops because of D-region absorption and lossy ground reflections. On the higher frequency bands where horizontal beam (Yagi) antennas are used, this means towers exceeding 60 feet — or even better, more than 100 feet. At the lower frequencies (below 10 MHz) vertical antennas on unobstructed, treeless sites, situated over moist earth and equipped with sufficiently large ground systems are needed to obtain 5- to 10-degree take-off angles. These are the antenna systems DXers dream about and a few build.

To obtain the highest probability of "reflecting" from an E_S "cloud," a fairly wide beamwidth should be used. Because the beamwidth of Yagis (50 to 60 degrees) is greater than the beamwidth of rhombics (20 to 30 degrees), Yagis are preferred for "hitting the clouds." E_S clouds usually measure about 6 by 60 miles (10 by 100 km) in length and about 1 mile (1.6 km) in thickness. They are oriented along the geomagnetic field line direction — i.e., roughly north-south over the United States. Their thin, dense configuration results in mirror-like *reflections* rather than the hundreds of kilometer thickness *refractions* that are characteristic of the F-region of the ionosphere. Signal strengths are, on the average, 25 dB higher due to the reflective as compared to refractive mode.

Another rule of thumb for E_S operation is to use the *lowest* frequency that still doesn't have excessive absorption. (For example, during daytime, don't use 10 meters if 20 is usable.) The probability of a good E_S opening occurring is greater with the lower frequency band.

There are trade-offs involved, however, in choosing either the lower or higher bands. You may, for example, want to select a lower band where E_S openings, though more probable, are limited by a higher take-off angle, more hops, and more loss. Or you might choose to go to a higher band for that occasional E_S opening, taking advantage of smaller take-off angles, fewer hops, and diminished losses (since the antenna is proportionately higher, with respect to wavelength, than the lower band version). Of course, if you're a 6-meter DXer, there's no choice — you'll have to wait for the E_S opening.

Early indications of E_S openings on the higher-frequency bands can be obtained by monitoring beacons on 6 and 10 meters and on CB channel 19. You can also monitor locally unused TV channels 2 through 5 for 6-meter openings. You may want to try W3ASK's system described in the June, 1985 "DX Forecaster." The lower frequency bands don't need beacon monitoring because E_S openings (sunrise and sunset hours) are available most of the time.

last-minute forecast

DX conditions are expected to be best for the higher frequency bands (10 to 30 meters, which are daytime bands) from the 12th through the 21st of the month, providing long and short skip openings. Look for short skip conditions the remainder of the month. The lower frequency bands (30 to 40 meters) should be best the first and the last weeks of the month, including daytime signals when the solar flux is below 70 units. The 80- and 160-meter bands will provide only fair DX conditions because of atmospheric noise build-up, except for some E_S short skip openings toward the end of the month.

The Aquarid meteor shower starts around the 18th, peaks about the 28th, and lasts until approximately August 7th. The maximum radio-echo rate will be approximately 34 per hour. The full moon is on June 29th, lunar perigee on the 21st, and summer solstice on the 21st at 1630 UT.

band-by-band summary

Six meters will provide occasional openings to South Africa and South America around local noontime via the E_S short-skip propagated mode.

Ten and fifteen meters will have many short-skip openings, and long skip during high solar flux to most southern areas of the world during daylight hours. Some enhanced trans-equatorial openings associated with disturbed ionospheric-geomagnetic conditions may occur in the evening hours this month.

Twenty, thirty, and forty meters will be useful for DX operation from most areas of the world during the daytime and into the evening almost every day, either long-skip to 2500 miles (4000 km) per hop or short-skip E_S to 1000 miles (1600 km) per hop. Since the period of daylight is now at its peak, high maximum usable fequencies will be practical for many hours of distance operation.

Thirty, forty, eighty and one-sixty meters will provide opportunity for nighttime DX operation. However, there will be many nights later this month when only 30 and 40 meters will be usable because of high thunderstorm QRN on 80 and 160. Nevertheless high signal levels, the result of short-skip E_S conditions, may help to overcome the static toward the end of the month.

GMT	PDT	WESTERN USA N	NE	E	SE	S	SW	W	NW	MDT	MID USA N	NE	E	SE	S	SW	W	NW	CDT	EDT	EASTERN USA N	NE	E	SE	S	SW	W	NW
0000	5:00	20	20	20	12	20	10	10	15	6:00	20	20	20	12	20	10	10	15	7:00	8:00	20	20	20	12	20	10	10	15
0100	6:00	20	*30*	*20*	15	20	10	10	15	7:00	20	20	*20*	15	20	10	10	15	8:00	9:00	20	20	*20*	15	20	10	10	15
0200	7:00	15	*30*	*20*	15	20	10	10	15	8:00	20	20	*20*	15	20	10	10	15	9:00	10:00	20	20	*20*	15	30	10	10	20
0300	8:00	15	*30*	*30*	20	20	10	10	15	9:00	20	*30*	*30*	20	30	10	10	20	10:00	11:00	20	*30*	*30*	20	30	10	10	20
0400	9:00	15	*30*	*30*	20	30	10	10	15	10:00	15	*30*	*30*	20	30	10	10	20	11:00	12:00	20	*30*	*30*	20	30	10	10	20
0500	10:00	20	20	20	20	30	10	10	15	11:00	20*	*20*	20	20	30	10	10	20	12:00	1:00	*30*	*20*	20	20	40	12	12	20
0600	11:00	20	20	15	20	30	10	12	15	12:00	20	*20*	20	20	30	12	12	20	1:00	2:00	*30*	*20*	20	*20*	40	15	15	*20*
0700	12:00	20	20	15	*20*	30	12	15	20	1:00	20	20	20	*20*	40	15	15	*20*	2:00	3:00	*30*	*20*	20	*20*	40	15	20	*20*
0800	1:00	20	20	20	*20*	40	15	20	20	2:00	*30*	20	15	*20*	40	15	20	*20*	3:00	4:00	20	20	20	*20*	*40*	20	20	*30*
0900	2:00	20	*20*	20	*30*	40	15	20	20	3:00	*30*	*20*	15	*30**	*40*	20	20	*30*	4:00	5:00	20	20	20	*30*	*40*	20	20	*30*
1000	3:00	*30*	*20*	20	*30*	*40*	20	20	20	4:00	*30*	*20*	15	*20*	*40*	20	20	*30*	5:00	6:00	20	20	15	20	*40*	20	20	20
1100	4:00	*30*	20	20	20	*40*	20	20	20	5:00	20	20	15	20	*40*	20	20	*30*	6:00	7:00	20	20	15	20	*40*	20	20	20
1200	5:00	*30**	20	15	20	*40*	*20*	20	*20*	6:00	20	20	12	20	*40*	*20*	20	20	7:00	8:00	20*	20	12	15	*40*	*20*	30	20
1300	6:00	20	20	12	20	*40*	*20*	*20*	*20*	7:00	20	20	12	15	*40*	*20*	*30*	20	8:00	9:00	15	20	12	12	*40*	*20*	30	20
1400	7:00	20	20*	12	15	*40*	*20*	*30*	*20*	8:00	20	20	10	12	*40*	*20*	*30*	20	9:00	10:00	15	20	10	12	*40*	*20*	*30*	20
1500	8:00	20	15	10	12	*40*	*20*	*30*	*30*	9:00	20	20	10	12	*40*	*20*	*30*	20	10:00	11:00	20	20	10	12	*40*	*20**	*30*	20
1600	9:00	20	15	10	12	*40*	*20*	*30*	*30*	10:00	20	15	10	12	*40*	*20*	*30*	20	11:00	12:00	20	15	10	10	30	*20*	*30*	20
1700	10:00	20	15	10	12	30	20	*30*	*30*	11:00	20	15	10	10	30	20	*30*	20	12:00	1:00	20	15	10	10	20	20*	*30*	20
1800	11:00	20	15	12	10	20	15	*30*	20	12:00	20	15	12	10	20	15	*30*	20	1:00	2:00	20	15	12	10	20	12	*30*	20
1900	12:00	*30*	15	15	10	20	12	20	20	1:00	*30*	15	15	10	20	12	15	20	2:00	3:00	20	15	15	10	20	12	15	20
2000	1:00	*30*	20	15	10	20	10	12	20	2:00	*30*	20	15	10	20	10	12	20	3:00	4:00	*30*	20	15	10	20	10	12	20
2100	2:00	20	20	20	10	20	10	10	20	3:00	*30*	20	20	10	20	10	10	20	4:00	5:00	*30*	20	20	10	20	10	10	20
2200	3:00	20	20	20	10	20	10	10	20	4:00	20	20	20	10	20	10	10	15	5:00	6:00	*30*	20	20	10	20	10	10	15
2300	4:00	20	20	20	10	20	10	10	20	5:00	20	20	20	10	20	10	10	15	6:00	7:00	20	20	20	10	20	10	10	15
JUNE		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN			ASIA FAR EAST	EUROPE	S. AFRICA	CARIBBEAN S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.

*Look at next higher band for possible openings.

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WANTED: Anything made by Kinemetrics True Time, Inc. Receivers, WWV clocks, time code readers, manuals, incomplete assemblies, etc. C.K.Barber, 3112 Ursula Street, Aurora, CO 80011.

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WANTED: Operations manual Hammarlund 105TR. Andy Nagy, 1444 Dill Road, South Euclid, Ohio 44121.

YAESU OWNERS: Hundreds of modifications and improvements for your rig. Select the best from fourteen years of genuine top-rated Fox-Tango Newsletters by using our new Comprehensive Index. Only $4 postpaid with Rebate Certificate creditable toward Newsletters purchases. Includes famous Fox-Tango Greensheet and Filter information for your rig (if specified). N4ML, FTC, Box 15944, W. Palm Beach, FL 33416. Telephone (305) 683-9587.

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ATTENTION AMATEURS Send for Free discount catalog. Amateur Communications, 2317 Vance Jackson, San Antonio, TX 78213. (513) 734-7793.

BEAM HEADING CHART, 10 page report in binder with 9 data fields calculated from your exact QTH to over 540 DX locations. $9.95 from John Daley, KB6JGH, PO Box 4794, San Jose, CA 95150.

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ANNOUNCING Iowa's new Computer BBS. (319) 557-9659. 24 hours per day. Contact User -8, AI9D for more information.

IBM/APPLE COMPUTER full feature log program, "Hamlog". Send SASE KA1AWH, Box 2015, Peabody, MA 01960.

TRANSFORMERS wound to 300 watts. Peter Robson Co., 18 Washington Trail, Hopatcong, NJ 07843.

ELECTRON TUBES: Radio, TV & Industrial Types—Huge inventory. Send for 80% off tube listing. Call Toll Free (800) 221-5802 or write Box HR, Transleteronic, Inc., 1365 39th Street, Brooklyn, NY 11218 (718) 633-2800.

THE GOOD SAM HAMS invite RV operators to check in the Good Sam Ham net 14.240 *Sundays 1900Z also 3.880 Tuesdays at 2359Z. Net control N5BDN, Clarksville, IN*

OLD RADIO transcription discs wanted. Any size, speed. W7FIZ, Box 724 HR, Redmond, WA 98073-0724.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $10.00. Beginners RTTY Handbook $8.00. PO Box RY, Cardiff, CA 92007.

IMRA, International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, New York 10538.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. Daily Electronics, PO Box 5029, Compton, CA 09224. (213) 774-1255.

CUSTOM MADE EMBROIDERED PATCHES. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept 301, 4202 N. Drake, Chicago, IL 60618.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

PRINTED CIRCUIT BOARDS and kits for QST articles. Call or write for information. A&A Engineering, 7970 Orchid Drive, Buena Park, CA 90620. (714) 521-4160.

PACKET/ASCII/BAUDOT/CW for IBM-PC. SASE to: Emile Alline, 773 Rosa, Metairie, LA 70005.

CABLE TV CONVERTERS/DESCRAMBLERS. Guaranteed lowest prices in US. Jerrold, Hamlin, Zenith -- Many others. Lowest dealer prices! Orders shipped within 24 hours! Mastercard, VISA, COD accepted. Free Catalog—call (80) 345-8927 only. Pacific Cable Co., Inc., 7325-1/2 Reseda Blvd,#1012, Reseda, CA 91335. (818) 716-5914.

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2 METER AMP KITS: 8877 legal limit kit $395. 3CX800A7 900W kit $325. Also HV power supplies, CX600N relays, parts and EME newsletter. SASE for catalog. 2 Meter EME Bulletin, 417 Staudaher St., Bozeman, MT 59715.

COMING EVENTS

Activities — "Places to go . . ."

CALIFORNIA: Flea Market/Boneyard Sale. Foothill College, Los Altos Hills. March-Sept. 2nd Saturday of every month 7 AM Sellers. 7:15 AM buyers. Talk in 145.27(-) or 147.570 simplex. FCC exams (408) 255-9000.

WASHINGTON: Central Washington Hamfest sponsored by the Apple City Amateur Radio Club, W7TD. June 14 and 15, Rocky Reach Dam near Wenatchee. Registration $5 for Amateurs. Others $1.00. Under 12 free. Banquet $7.00. Equipment displays, Swap Shop, ARRL VE license exams. Free camper/trailer space with power provided by Park. Talk in on 146.07/67 or 146.49 simplex.

NEW YORK: The Mt. Beacon Amateur Radio Club Hamfest, Saturday, 12, Arlington Senior High School, Poughkeepsie/LaGrange. Tickets $3.00 (hams and computer hobbyists). Tailgating space $4.00. Tables $6.00—one free table and admission. Doors open 8 AM. Talk in on 146.37/97and 146.52. For more info: Julius Jones, W2IHY, RR 2, Vanessa Lane, Staatsburg, NY 12580 (914) 889-4933.

INDIANA: The Lake County Amateur Radio Club's 14th annual Fathers' Day Hamfest, Sunday, June 15, Lake County Fairgrounds Industrial Building. Free parking. Tables available. Admission $3.00. No advance. 8 AM to 2 PM. Setup 6 AM. MARS, ARRL, ARES information. Refreshments. Overnight accommodations nearby. Talk in on Lake County ARC repeater 147.84/24 or 146.52 simplex. For further information: Bill DeGeer, W9TY, 3601 Tyler Street, Gary, IN 46408 (219) 887-5413 evenings after 6 PM.

NEW HAMPSHIRE: Fly-in to New Hampshire's 2nd largest Amateur Radio/electronic flea market, Saturday, June 21, Manchester Municipal Airport. Sponsored by the New Hampshire FM Association. Rain date Sunday, June 22. Starts 9 AM. General admission $1.00. Sellers $5.00, bring own tables. Commercial displays welcome. Refreshments available. Preregistration to 123 Woodlawn Circle, Portsmouth, NH 03801. Talk in on 146.52 FM. For information: Doug Aiken, K1WPM, (603) 622-0831 or Pete Henriksen, WA1RCF, 123 Woodlawn Circle, Portsmouth, NH 03801. (603) 431-5432.

WISCONSIN: The South Milwaukee Amateur Radio Club's annual Swapfest, Saturday, July 12. American Legion Post 434, 9327 South Shepard Avenue, Oak Creek. 7 AM to 4 PM. Parking, picnicking, food and refreshments available. Admission $3.00 includes "happy time" with free beverages. The Milwaukee Volunteer Core Group will conduct Amateur Radio exams during the day. For more details, including a map, write: South Milwaukee ARC, PO Box 102, South Milwaukee, WI 53172-0102.

1986 "BLOSSOMLAND BLAST" Sunday, October 5, 1986. Write "BLAST", PO Box 175, St. Joseph, MI 49085.

CALIFORNIA: FCC exams, Novice-Extra. Sunnyvale VEC ARC. (408) 255-9000 24 hour. 73, Gordon, W6NLG, VEC

PENNSYLVANIA (Meadville) -- July 5. Firecracker Hamfest by Crawford Amateur Radio Society. Electronics, radios, computers. New location at Meadville Recreation Center just off Rt. 27 E in Meadville. All indoors; no rain out. One olympic and two child-size deluxe swim pools with lockers and showers and other sports available in Center. Free indoor fleamarket space; bring table. Dealers: free indoor space and table(s) available by preregistration; reserve early. Admission adult $2.00, children free, swimming for small extra. Talk in 144.53/145.13, W3MIE/R. For information write CARS Hamfest-86, PO Box 653, Meadville, PA 16335 or call Ben Ferer, KF3F, 814/724-2432.

KENTUCKY: The Antique Radio Club of America (ARCA) will hold a convention, June 11 through 14 in Louisville. Club members collect and restore antique wireless and radio equipment from before World War I through the 40's and 50's. At this year's convention there will be speakers, tours of radio collections and also tours of the Louisville area. Massive flea market and auction. Louisville offers much in the way of recreation for the entire family. For information: ARCA, 81 Steeplechase Road, Devon, PA 19333 or call (215) 688-2976.

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NEW TITLES

FIRST STEPS IN RADIO

by Doug DeMaw, W1FB

This new anthology has been taken from DeMaw's '84 and 85 series in QST magazine. It has been written to give beginners the basic electronic theory needed for upgrading and the ins and outs of how radios work. Using a building block approach, DeMaw first explains what the different componants are, then assembles them into basic circuits and ends up with how these circuits work in your radio. You also get articles on antennas, propagation and beginners level RFI problems and suggestions on how to resolve them. Great review for more experienced Hams. Perfect for the beginner. © 1985.

☐**AR-FS** **Softbound $4.95**

LANDMOBILE AND MARINE RADIO TECHNICAL HANDBOOK

by Edward Noll, W3FQJ

This is THE HANDBOOK for those who operate, install or service two-way radios. Covers private landmobile services, marine radiotelephone and radiotelegraph, marine navigation and Citizen's Band. An excellent reference book for those studying to pass the NABER technician certification exam. Areas covered include: transmission characteristics and modulation systems, basic solidstate theory, digital and microprocessor electronics, antenna systems, test equipment, repeaters and much more. © 1985 1st edition 576 pages.

☐**22427** **$24.95**

THE COMMODORE HAM'S COMPANION **by Jim Grubbs, K9EI**

Here's your guide to using the Commodore C-64 computer in your Ham shack. Good solid information on where to find software and hardware for CW, RTTY, AMTOR, SSTV, propagation prognostication, antenna modeling, satellite tracking and much more. Includes a list of over 80 sources of software and hardware. Also includes a bibliography of over 60 magazine articles and reviews about using the Commodore. ©1985, 160 pages, 1st edition.

☐**JG-CC** **Softbound $15.95**

Please add $3.50 to cover postage and handling.

Ham Radio's Bookstore

Greenville, NH 03048

ILLINOIS: The Egyptian Radio Club will hold its 57th annual Hamfest, Sunday, June 8, 8 AM to 3 PM, at their clubhouse. Free 10' flea market spaces available on first come basis. Additional spaces $5.00. Tickets $1.00 advance, $2.00 each or 3/$5.00 at the door. Talk in on 146.16/76 or 146.52 simplex. For information or tickets SASE to Egyptian Radio Club, PO Box 562, Granite City, IL 62040.

BRITISH COLUMBIA: Maple Ridge Hamfest, July 12 and 13, St. Patricks Center, 22589 121 Avenue, Maple Ridge. Admission Hams $6.00; non-hams over 12 $3.00. Under 12 free. Two hams in family $9.00. Commercial displays, flea market, food, children's programs. Nearby shopping and recreation areas. Camper space, no hookups. Talk in on 146.20/80 and 146.34/94. For more information: Bob Haughton, VE7BZH, Box 292 Maple Ridge, BC V2X 7G2 or phone (604) 467-4915.

NEW YORK: Ham Family Day, June 1, Becks Grove, Blosvale. Sponsored by the Rome Radio Club. QRP contest, flea market, tech talks. Non-ham activities. Talk in on 146.28/88 and 146/34/94. For information: Rome Radio Club, Box 721, Rome, NY 13440 or call William Effland (315) 853-5700.

MINNESOTA: The North Area Repeater Association's Swapfest and exposition, Saturday and Sunday, June 7 and 8, Minnesota State Fairgrounds in St. Paul. Free overnight parking for SC campers June 6 and 7. Exhibits, commercial dealers, giant outdoor flea market. Amateur license exams will be given. Admission $4 advance, $5 at gate. For more information, tickets, etc: Amateur Fair, PO Box 857, Hopkins, MN 55343 or call (612) 566-4000.

NEW JERSEY: The Raritan Valley Radio Club's 15th annual Hamfest, Saturday, June 21, Columbia Park, Denellen. Gates open 8 AM. Sellers spots $5.00/space or $10.00 multiple spaces. No tables supplied. Lookers $3.00 donation. Spouse and kids free. Food and drink available. Talk in on Club repeater W2QW/R 146.025/.625 and 146.52 simplex. For information call Dave, KA2TSM (201) 763-4849 or Bill, N2AZX (201) 467-7342, 8 AM to 5 PM.

PENNSYLVANIA: Harrisburg annual Firecracker Hamfest, Friday, July 4, Bressler Fire Company picnic grounds. Sponsored by the Harrisburg RAC. Tailgating no charge with admission of $3.00. XYL and kids free. Nearby restaurants and motels. Parking on grounds. Talk in on 52-52 simplex or local repeaters. For details/table info: Dave, KC3MG, 131 Livingston St, Swatara, PA 17113 or call (717) 939-4957.

WEST VIRGINIA: The 8th annual TSRAC Wheeling Hamfest/Computer Fair, Sunday, July 20, Wheeling Park. 9 AM to 4 PM. WV's largest. Dealers welcome. 30,000 sq. ft under roof. 5 acres of flea market. Family activities at Park. Admission $3.00 *in advance, $4.00 at door. To reserve space contact:* Jay Paulovicks, KD8GL, RD 3 Box 238, Wheeling, WV 26003 (304) 232-6796. For tickets, TSRAC, Box 240, RO 1, Adena, OH 43901 (614) 546-3930.

HAM DATA
C-64 Software

SUPER LOG

Super log gives you all the advantages of a computerized data base without significantly changing the traditional log format. For contesters, Super Log can be configured to either manually or automatically enter contact number as well as time of contact. Make an error and you can easily go back and edit the entry. Super Log also allows you to print out either selected contents or the whole log. Will print QSLs.

☐**HD-SL** (For C-64) **$19.95**

CONTEST LOG

This disk contains four different contest programs; ARRL Sweepstakes, Field Day, Universal WW Contest log, plus a dupe checking routine. Each program is designed for real time use. It automatically enters date, time, band and serial number for each contact. A 24-hour clock is displayed at the top of the VDT screen. When the contest is over, the program will print your results listing all duped and scored contacts in serial sequence with all the necessary information as well as completed score at the bottom of the page.

☐**HD-CL** (For C-64) **$24.95**

MASTER LOG

Over three years of development went into this program. It creates a file of 2100 individual records with up to 13 different entries per record. Master Log can do a search and select based upon time, frequency, mode or any of the other variable parameters. It keeps track of DXCC and WAS status, prints QSL labels and can search its whole file in less than 5 seconds! Complete documentation is included to help you learn and use this truely state-of-the-art logging program.

☐**HD-ML** (For C-64) **$28.95**

Please enclose $3.50 for shipping

HAM RADIO'S BOOKSTORE

Greenville, NH 03048

MARYLAND: The Frederick Amateur Radio Club's 9th annual Hamfest, June 15, Frederick Fairgrounds. 8 AM to 4 PM. Admission $3.00. Tailgaters $2.00 extra. YL's and kids free. Gates open for exhibitors 8 PM June 14. Exhibitors tables $10.00 first, $5.00 each extra. For information: Jim Kasunic, KA3LPC, 9419 Highlander Ct, Walkersville, MD 21793.

INDIANA: The 16th annual Indiana ARRL Convention and Indianapolis Hamfest, Saturday and Sunday, July 12 and 13, Marion County Fairgrounds, Indianapolis. Featuring the largest electronic flea market and new Amateur Radio equipment display in the state. Gate fees $5.00 with free parking on grounds. Gates open 6 AM. Large covered flea market. Inside tables $8 each. 2m Fox Hunt, homebrew contest. QSL card contest. Free camper hookups available on grounds. Nearby restaurants and motels. For information contact Bill Evans, WB9BEN (317) 745-6389

MASSACHUSETTS: The MIT UHF Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday, June 18, 7 PM, MIT Room 1-134, 77 Mass Avenue, Cambridge MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-0160/646-1641 or Craige Rodgers (617) 494-1986. Exam fee $4.25. Bring copy of current license (if any), two forms of picture ID and completed form 610 available from FCC in Boston (223- 6609)

PENNSYLVANIA: The 12th annual Warminster Amateur Radio Club's Hamfest, Sunday, June 8, Middletown Grange Fairgrounds, Penns Park Road, Wrightstown. Gates open 7 AM (vendors 6 AM). Donation $3.00. XYL's and kids free. Outdoor tailgating $5.00 per space. Indoor space $5.00 with 8' table and power available. Talk in on 147.69/09 and 146.52. For information and preregistration: Chuck Dunn, KA3FQQ, 1414 Bradley Lane, Warminster, PA 18974.

OPERATING EVENTS
"Things to do . . ."

The LINCOLN TRAIL ARC will operate W4BEJ from Abraham Lincoln Birthplace National Historic Site near Hodgenville Kentucky during Field Day June 28 and 29. For a commerative certificate send QSL and SASE to LTARC, PO Box 342, Vine Grove, KY 40175.

ACB RADIO AMATEURS a special-interest affiliate of the American Council of the Blind (ACB) will operate special event station KW4U from 0000Z June 29 to 2400Z July 5 from the Hilton Hotel in Knoxville, TN in conjunction with ACB's silver anniversary convention. Submit QSL card confirming QSO with KW4U during convention week and receive an attractive commemorative certificate. Send QSL's to John McCann, KW4U, 2105 N. Illinois Street, Arlington, VA 22205.

BUZZARD ROOST REPEATER will operate special event station WB0UPK,1500Z July 4to 0100Z July 5, from the Neligh Mills, a state historical site. QSL's may be obtained by sending SASE to WB0UPK, 804 East 1st Street, Neligh, NE 68756.

EASTERN CONNECTICUT ARA will operate special event station, K1MUJ, to help celebrate the Woodstock Connecticut Tercentenary on July 4. For a commemorative certificate send SASE and QSL card to KB1SP.

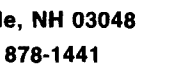
ARGONNE ARC will operate W9QVE, June 21 from 11 AM to 4 PM CST to commemorate the 40th anniversary of the establishment of the US National Laboratory System of which, Argonne is the first. Send QSL and SASE to AARC, PO Box 275, Argonne, IL 60439.

SOUTH JERSEY RADIO ASSOCIATION, the oldest radio club in the US, will operate a special event station, K2AA, from 1200Z June 7 to 0400Z June 16 to celebrate its 70th birthday. For a commemorative QSL send SASE and QSL or log info to South Jersey RA, PO Box 1026, Haddonfield, NJ 08033.

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HAM RADIO MAGAZINE

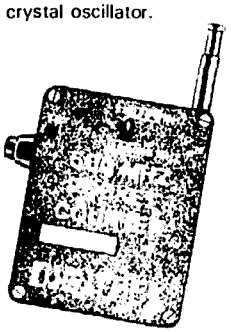
Greenville, NH 03048

(603) 878-1441

product REVIEW

500 MHz Pocket Counter

Here's a neat little product every ham will want to have: a 500-MHz counter from Digitrex that fits into a shirt pocket, measures just 1 x 3.5 x 2.7 inches, and weighs less than 8 ounces *with* the 9-volt battery installed. The LED readout display measures 0.14 inch; the unit draws less than 100 mA. Battery life is estimated to be several weeks in normal use. The unit uses six ICs, five transistors, three diodes, and a 2.56 MHz internal crystal oscillator.



The counter has a sensitivity of less than 50 millivolts, 1-500 MHz, 1.5 volts RMS maximum and the resolution is four digits (switchable to six) down to 1 kHz. A seven-digit modification that gives resolution down to 100 Hz is available. The counter has a short-term accuracy of 1 ppm when calibrated against a known signal using an internal trimmer capacitor. Over the long term, its accuracy is 1 kHz at 500 MHz over a temperature range of 10-40 degrees C.

To use the unit, you first install a 9-volt transistor battery, extend the internal 19-inch antenna, select the MHz range, and depress the counter's on/off button, which also functions as a press-to-read control. All you do is read the frequency — it's really as simple as that!

Not having a lab-grade frequency generator with which to compare readings limits the scope of this review to using the VFOs (and internal frequency calibrators) of several different transmitters as frequency standards. All readings were found to be within manufacturer's specifications.

This is really a "neat" product and at the price, it's one good value for your dollar. While you can't expect lab-grade measurements, you'll be pleased with the unit's accuracy.

The price is $49.95 plus $2 shipping and handling. For more information, contact: Digitrex, 1005 Bloomer, Rochester, Michigan 48063.

— N1ACH

NEW products

triple output power supply

MFJ Enterprises, Inc. has announced the release of its Triple Output Lab Power Supply. For $149.95 (plus $6.00 shipping and handling), the MFJ-4002 offers two variable 1.5 to 20 VDC outputs at 0.5 amp and one fixed 5 VDC output at 1 amp. It's designed for heavy duty commercial use to give years of trouble-free service.

There's plenty of voltage and current for both analog and digital circuits. It's ideal for education, circuit design, product development, testing and repair, quality control and production.

Separate transformers are used for completely isolated outputs. This allows the outputs to be connected in series or parallel for higher voltage or current. It's short circuit protected, has excellent line regulation (typically 0.1 percent V) and load regulation (typically 0.1 percent) and has very low ripple.

Two lighted 3-inch precision meters are provided for monitoring voltage and current simultaneously. Rugged 5-way binding posts are used for all outputs. A separate binding post is used for chassis ground.

The MFJ-4002 is made in America and built with heavy guage aluminum to take lots of abuse. It measures 12x36x6 inches, uses 110 VAC with a 3-wire safety power cord and fast acting pop-out fuse. For details, contact MFJ Enterprises, Inc., 921 Louisville Road, Starkville, Mississippi 39759.

Circle #301 on Reader Service Card.

new heathkit catalog

A wide variety of electronic products in kit form and assembled versions are showcased in the new Heathkit Catalog from Heath Company of Benton Harbor, Michigan. A new 2-meter handheld/mobile system featuring greater ease of operation and greater flexibility is offered, as is a low-cost, user-installed satellite sytem featuring the General Satellite Dish and the Norsat Satellite Receiver.

The HW-6502 2-Meter Hand-held Transceiver comes complete with a built-in CTCSS encoder, S/BATT meter, squelch control and thumbwheel channel selector switches with ±5 kHz switch. The operating frequency range is 144 MHz to 147.995 MHz and sensitivity is 0.25μV. The mobile console unit, HWA-6502-2, houses the transceiver and is able to provide 25-watt mobile operation with the optional 25-watt 2-meter amplifier. The console, which includes power, UHF and VHF switches and lit frequency display, measures 2-1/4 x 6-1/2 x 8-1/2 inches (excluding projection).

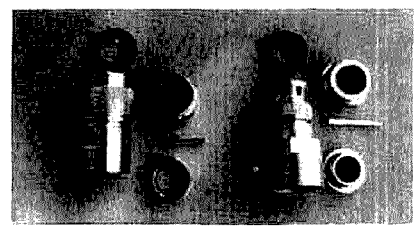
The Heathkit Satellite System combines the Norsat receiver and the General Satellite Dish for the user to install and receive satellite relayed programming. The Heathkit Site Survey kit is used to determine acceptable sites for positioning the system's dish antenna.

Over 400 electronic products are offered in the new Heathkit Catalog. To receive a free copy, contact Heath Company, Benton Harbor, Michigan 49022.

Circle #302 on Reader Service Card.

cable connectors

Nemal Electronics International has introduced a new line of connectors designed to fit the Belden 9913- and 8214-type cables. The new connectors are available from stock in both type N (part #NE720) and BNC (part #NE860) series and will accommodate the 9-1/2 to 11 gauge center conductors in these and other similar cables.



Both series of connectors meet the electrical and mechanical requirements of MIL-C-39012 and incorporate silver plated contacts and teflon insulation. Each connector is fully compatible with all other standard connectors in its series.

For information, contact Nemal Electronics International, 12240 NE 14th Avenue, North Miami, Florida 33161.

Circle #303 on Reader Service Card.

new VOM

Mercer Electronics, a division of Simpson Electric Company, has introduced a new VOM; Model 9120 has 25 ranges including dB, 20,000 ohms/volt DC sensitivity (5,000 ohms/volt AC) and a frequency response up to 100 kHz on 3, 12, and 60-volt AC ranges. It will measure up to 12 Amperes DC, has a 3-volt AC range, and provides excellent resolution. DC accuracy is ±3 percent fs. Convenience features include a front panel polarity reversal switch, single-knob range/function switch with an "off" position, an output jack for DC isolation, and a large, easy-to-read, 3.5- inch mirrored, color-coded scale.

Also included is a high-energy fusing system along with standard fusing and diode meter protection. The unit measures 4 x 6 x 1-3/4 inches and weighs 13 ounces. It comes complete with batteries, test leads and an operator's manual.

For details, contact Mercer Electronics, 859 Dundee Avenue, Elgin, Illinois 60120.

Circle #304 on Reader Service Card.

THE GUERRI REPORT

by Ernie Guerri, W6MGI

new communications satellites will use laser links

A new concept in communications satellites is in the development stages in European laboratories. The large data capacity of optical communications links will be exploited by incorporating a carbon dioxide laser into a geosynchronous satellite. The CO_2 laser operates in the far infrared region, at a wavelength of about 11 microns (27,000,000 MHz).

In spite of substantial technical hurdles, its benefits are thought to justify the effort. Advantages include simultaneous bi-directional data capacity of more than 1 gigabit/sec., with an accuracy of about 10 bits (an error rate of 1 part in 10^9) at an altitude of about 37,000 km. The transmissions would have high resistance to interference and jamming because the frequencies are removed from present high-density traffic, and the laser beam is very narrow — less than 1 km wide at geosynchronous altitude.

The choice of the CO_2 laser is based on the fact that highly sensitive detectors already exist, the technology for the needed power levels is well advanced, and relatively conventional optics can be used. The system would consume about 250 watts in order to produce a little more than 1 watt of modulated power. Transmitter bandwidth will be about 5 GHz, permitting higher data rates as technology and traffic demand mature. Two technical areas will need considerable attention in order to achieve successful operation. The optical receivers need to be cooled to −173 degrees C (100 degrees K) for optimum sensistivity, and the pointing and tracking system must be able to keep the very narrow beam centered on the respective receivers. Present technology is marginal in each of these areas.

Components for the planned satellite will begin to be assembled as a laboratory breadboard as early as 1987. The first flight tests are planned for 1990-91, and an operational satellite is contemplated for 1994-95.

emergency transmitter gives directions when lost

The Federal Aviation Administration requires that all aircraft carry an emergency locator transmitter (ELT) that can help pinpoint the aircraft's location in the event of a crash. The device is a small battery-powered unit that transmits on 121.5 and 243.0 MHz. A newly-developed unit takes full advantage of today's IC's to greatly enhance the data sent by the beacon.

As the aircraft flies, a circuit in the locator continuously enters the current position of the plane from the aircraft's LORAN-C navigation system. In the event of a crash, the locator broadcasts a synthesized speech message beginning with the word "mayday" and stating the aircraft's registration number, its longitude/latitude, the elapsed time since impact (very important in harsh environments), and other key information — all in plain English. This improvement will simplify search and rescue operations and expand the number and type of observers who can participate in radiolocation searches — Amateurs, SWLs, scanner buffs, and so on.

chip inductors join new component ranks

Since the beginning of the era of solid-state circuits, component manufacturers have been quick to keep pace with designers' needs. Components of all types have shrunk in size and taken on the shapes appropriate to the application. Inductors have resisted most attempts at going "monolithic" and have caused a lot of otherwise elegant designs to be much larger and less reliable than the designer might have wished.

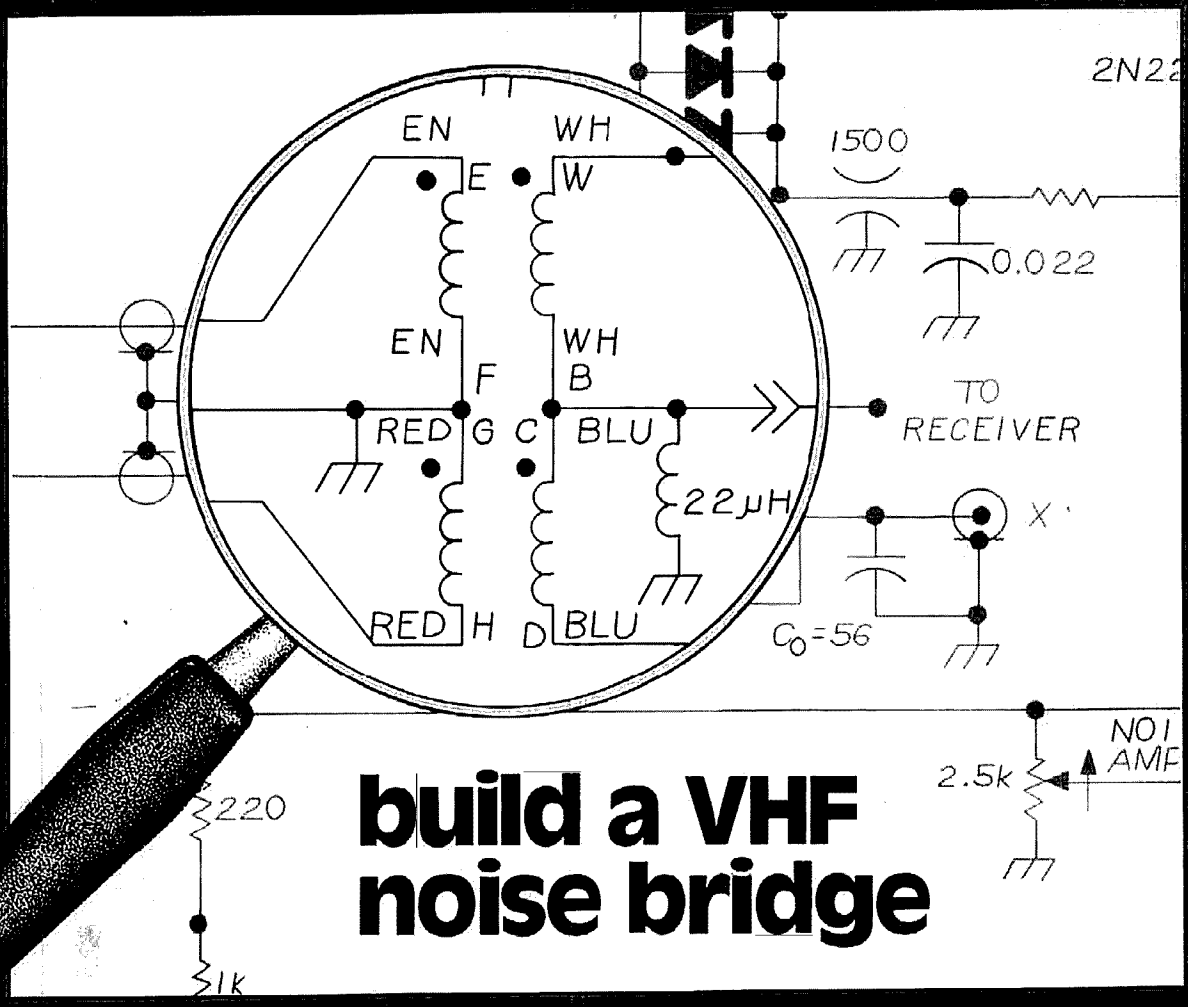
We still haven't found any magic way to solve the problem, but there has been significant progress in magnetic materials combined with creative new packaging techniques. Given incentive by the move to surface-mounted components, leadless "chip" inductors are making their debut. Using ceramic or ferrite cores, these ultra-miniature inductors can be made with very small gauge wire because of the low power levels associated with modern IC circuits. Eliminating leads also reduces stray capacitance and parasitic inductance, thereby making the component a little closer to ideal.

Chip inductors have evolved in a few "standard" packages, which makes the layout job easier for the circuit designer. Inductance values cover the full range from about 3 nH to 1 mH, and include both fixed and variable type. Because these inductors mount directly on the circuit substrate, great care must be taken in the choice of temperature coefficients. If the inductor is not matched to the substrate, expansion and contraction can cause severe lead stress, and in the worst cases can actually lift the inductor from the substrate. Shielding poses serious problems on all but small inductance values. At present, shielding simply takes up real estate — which of course defeats the purpose of using the chip inductors. Notwithstanding the difficulties, before long you can expect to be staring at a totally unfamiliar-looking component on a circuit board — a "chip" inductor.

ham radio

ham radio

ANNUAL VHF/UHF SPECIAL ISSUE



build a VHF noise bridge

ham radio magazine

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T. H. Tenney, Jr., W1NLB
publisher

Rich Rosen, K2RR
editor-in-chief
and associate publisher

Dorothy Rosa, KA1LBO
assistant editor

Joseph J. Schroeder, W9JUV
Alfred Wilson, W6NIF
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editorial production

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assistant publisher

Rally Dennis, KA1JWF
director of advertising sales

Dorothy Sargent, KA1ZK
advertising production manager

Susan Shorrock
circulation manager

Therese Bourgault
circulation

cover art:
Barbara Smullen

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REFLECTIONS

the phone call

It took a late-night, long-distance phone call to remind me of what I used to easily perceive as the excitement of ham radio. The call came from a good friend who'd gotten into Amateur Radio at about the same time as I did, approximately 28 years ago. Catching up on recent history, I couldn't help noticing not only his present and continuing enthusiasm for the hobby, but also his accomplishments over the years. His interests included, but weren't limited to, antenna and propagation experimentation, modification of commercial equipment, the design and construction of new equipment, operation on the newer specialized communications modes, as well as teaching and helping other Amateurs. But what impressed me the most was his intensity — his desire to do the very best he could while enjoying what he was doing.

Not long ago I found myself feeling discouraged about the future of Amateur Radio. It's not difficult to get into this mood. Just listen to discussions on the air, at hamfests and at other meetings. . . or open up to the editorial page of any ham magazine. There you'll no doubt hear, or read, that the average age of Radio Amateurs in this country is rising steadily. If you really want to get discouraged, tune any of the HF bands (and the most popular VHF band). What do you hear?

- *"CQ DX, CQ DX."* (Translated, this means "I want to contact you, the rare station, as fast as I can and get your QSL card and then goodbye!")
- Or *"Hey, Goofball, this is my frequency and I'm not moving!"* (Translation: "I've been on this frequency for the past three hours and I own it.")
- Or *"Hey, Joe, are you sure I'm only 37.5 dB stronger than that other W1? By my figuring I should be at least 40 dB up from him!"* (Somewhere along the line many of us forgot that the FCC didn't intend for us to turn Amateur Radio into a horsepower race.)
- Or, as heard on 2 meters not too long ago, *"Tom, I just got this new Loudenboomer amplifier. It's wired for 110 volts. How do ya wire it for 220?"* (I guess that question wasn't on his Advanced class exam).

Is this what Amateur Radio is all about?

There are those who say we need some stimulation and that the solution is obvious: get more youth involved. Presto! Amateur Radio — if not the world itself — is saved. Quite frankly, I don't believe in simplistic cures and single-answer solutions. But it is true that in order for any organization to self-perpetuate, a constant influx of "newness" is required, be it youth itself or just youthful energy and spirit.

Even if we manage to attract young people to our hobby, it will still take Amateurs (like my friend) willing to share their enthusiasm and knowledge with them. By helping young people pursue their own interests, these Amateurs will encourage young people not only to enter Amateur Radio, but stay there. It won't be just young people who "save" Amateur Radio, but those hams like my friend who make that special effort.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

SEVERE SANCTIONS AGAINST CASUAL SWL'ING STILL REMAIN in the latest version of the "Electronic Communications Privacy Act of 1986" adopted unanimously by Rep. Kastenmeier's subcommittee May 14. Though the Act specifically exempts any station transmitting in an Amateur band, it is still based on the negative philosophy that a U.S. citizen has no right to tune to any radio transmission except as permitted by the government (see the editorial in February's Ham Radio). Furthermore, this latest version now defines "interception" of radio or other electronic communications as the interception of the transmission itself rather than its content! The Act provides for fines up to $10,000 and a year in jail for tuning in remote broadcast pickups, possibly any ship-to-shore communications, any kind of encoded transmissions, RCCs (older type car phones), and any FM subcarriars.

The Cellular Telephone Industry Didn't Get Exactly what it wanted; the penalty for receiving cellular would be only $500 and/or six months in jail. However, when Rep. Mike DeWine (R-Ohio) offered an amendment to limit the Act to encrypted communications or the disclosure of the contents of protected but unencrypted communications, Rep. Kastenmeier responded saying he wanted his bill to discourage casual listening -- even though the Justice Department has said it couldn't and wouldn't enforce the Act against casual listeners. That amendment, and another by DeWine to eliminate the six-month penalty for eavesdropping on cellular communications, were both voted down.

The Act (HR-3378) Now Goes To The House Judiciary Committee where it is expected to meet with little opposition. At the same time hearings on the Senate's version of the bill, S-1667, are expected to be scheduled in the very near future.

A 30-DAY WAIT BEFORE RETAKING AN AMATEUR EXAM IS INDEED UNNECESSARY, the FCC affirmed in deciding against ARRL's Petition for Reconsideration in PR Docket 85-31. The League had argued for retaining the delay because its program is set up so that ARRL VEs can use the same exam for some period of time, but the FCC said in its decision, "We will not retain it [the wait] merely to accommodate the administrative choices made by one or more VECs. It is far more important to eliminate unnecessary and outdated government regulations."

PRB-1 HAS APPARENTLY RESOLVED KA7TVC'S ANTENNA PROBLEM with Kirkland, Washington authorities (see June Presstop). His 2-meter vertical had been cited under a city ordinance restricting transmitting antennas, but Kirkland is considering its law firm's advice to revise its ordinance to agree with PRB-1 and has dropped its action against KA7TVC.

An Ordinance Limiting Antennas To Six Feet Above Roof Line in the District of Columbia was up for first hearing at presstime; ARRL has prepared a strong presentation against the proposal. Bills limiting all towers in New York State to 58 feet or tree-top height, whichever is lower, have been introduced in that state's legislature.

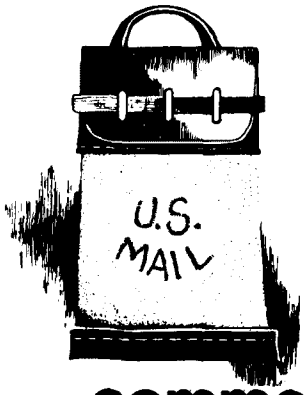
SUGGESTIONS AND COMMENTS FOR THE VEC PROGRAM ARE BEING SOUGHT by the Council for Amateur Radio Examining (CARE) in preparation for a meeting with the FCC tentatively set for August. The Washington meeting, for all accredited VECs, will review how well the program is working and how it might be improved. Write CARE, Box 688, Glanview, Illinois 60025.

A NEW BAND PLAN FOR 10-METER FM WAS SUPPORTED by nearly 100 Amateurs attending the 10-Meter FM Forum at the Dayton Hamvention. In brief it would change the repeater offset to 400 kHz instead of 100 kHz and use 29.50-29.68 MHz for the repeater outputs. Repeater inputs would then be 29.10-29.28, on 20 kHz centers. 29.3-29.4 MHz would be used for FM simplex, with 29.4-29.5 a "No-FM" slot for OSCAR downlinks so long as any Amateur 10-meter satellites remain operational. A formal Petition for Rule Making is being planned; Bob Heil, K9EID, would appreciate comments; write Bob at Box 78, Marissa, Illinois 62257.

Revision of 10-Meter Beacon Operation into a system similar to the 20-meter beacon system operating on 14.100 MHz has been proposed by the IARU. In Resolution B5-1 the Union would set aside 28.190-28.200 MHz for beacons, with a worldwide beacon network on 28.200 and regional networks from 28.190 to 28.199. Almost all 10-meter beacons now on the air operate between 28.200 and 28.300, but if the proposed expansion of 10-meter Novice privileges is adopted, the usefulness of these beacons -- most running low power -- would be seriously compromised. G3DME coordinates the International Beacon Project for IARU.

LABELS SPECIFYING RFI SUSCEPTIBILITY ON HOME ENTERTAINMENT EQUIPMENT are being sought by the ARRL in a Petition for Rule Making filed with the FCC. If adopted, the proposal would require every such device to carry an FCC-mandated label or tag that would specify just how susceptible that device is to interference from nearby transmitters.

In Canada VE3SR Has Decided To Fight The Judge's Decision that put him off the air as a "nuisance" to his neighbors' home entertainment equipment (see June Presstop). Though it now looks as if he may receive some support from commercial two-way users who are concerned about the possible impact of the decision on their operations, he'll still need help from the Amateur community to meet the estimated $15,000 cost of the appeal.



comments

amplifier parasitics

Dear HR:

Richard Measures's article in the April, 1986, issue, "Grounded-Grid Amplifier Parasitics," was most interesting. It provided insight into several amplifier circuit designs that are indeed troublesome.

I, too, experienced parasitic oscillations in my first SB-220. While the plate tuning capacitor did arc over, no damage was done. A local Amateur who was well versed in VHF equipment construction pointed to the silver mica capacitor and RF choke tied from each tube socket grid pin to ground as suspect. At his suggestion, the component leads were *reduced to* the shortest possible length. The choke was also re-positioned so that its "cold" end was laid against the chassis. This appeared to reduce the choke *Q* and lessen the probability of parasitic oscillation. After these simple modifications, the amplifier remained stable. The tuning capacitor never arc'ed again.

For the past eight years I've had first-hand production experience with several hundred HF linear amplifiers using the Eimac 8874 and 8877. To date I've not seen a single case of parasitic oscillation or instability of any kind with these tubes. In addition to careful circuit layout, the key to stable operation appears to be the use of Eimac's recommended tube sockets. It's no accident that a tube manufacturer specifies all relevant operating parameters — from airflow requirements, to element voltages, to such mundane items as heat-dissipating anode connectors and sockets. The E.F. Johnson socket specified for use with the 8874 can be mounted so that the grid pin terminal lugs are bent over and soldered directly to the socket saddle. The Eimac SK-2210 socket for the 8877 provides four short, direct, grid grounding spring clips. In both cases, the grid-to-ground connection has very low inductance and a large conductive area.

The possibility of interelectrode shorting demands the inclusion of a current limiting resistor in the plate supply. Without it, the tubes and HV rectifiers are in jeopardy. The resistance provides a voltage drop that's dependent upon the current drawn. The voltage drop can actuate a plate current relay that turns off the amplifier if the dreaded short occurs. This resistor, like the "feedback resistors" in AG6K's article, provides cheap insurance.

Reliable equipment operation isn't magical. Nor is it accidental. It should be *the heart* of the design.

Ray R. Heaton, NJ0G
Canon City, Colorado 81212

Dear HR:

Having had 36 — I repeat, *three dozen* — 572 tubes fail in a manner similar to that described by AG6K, in his April article (page 31), I found his conclusions very interesting and provocative.

The amplifier I was using was grid-driven, neutralized, and had parasitic suppressors in both plate and filament leads. The filament lead suppressors varied from several to as many as eight turns of No. 14 enameled wire wound on an 82-ohm 2-watt resistor. I had no arcing at the plate tuning capacitor to warn me because the capacitor is rated at 7000 volts. In every case, the big bang would result in one tube failing at a time. The tube would develop a grid-to-filament short and sometimes the filament would open.

I liked AG6K's explanation and cure, but several things keep sticking in my mind. He says the cure is 90 percent, not 100 percent effective. In my case, on three or four occasions I definitely saw the grid meter give a reverse kick just before the big bang. The tubes were operated Class C with high bias voltage.

He is right about lots of power being involved, because the bang would blow the 30-Ampere fuses in the 234-volt line feeding the power supply.

The trouble seems to have started around 1981. Tubes purchased before that time gave years of trouble-free operation. I have an old HXL-1 in which the tubes must be 15 years old. I may be wrong, but I came to the conclusion that the troubles began when the manufacturer changed the amplification factor from 160 to 200. (I have a feeling AG6K is going to write more on this.)

John Labaj, W2YW
Elsmere, New York 12054

directive antenna

Dear HR:

W4MB's article, "A New Class of Directive Antenna," (April, 1986, page 107) — with its coverage of the Landstorfer antenna — was very interesting. I was impressed when this idea was recently demonstrated by Frank Rutter, K3AW, at the Tropical Hamboree. Then I remembered that a little more gain can be developed with a straight shorter (5/4 wave) element Yagi which would be much simpler, mechanically, to reproduce. An example of this, the "extended double Zepp"-type Yagi beam antenna was described "The Extended Element Beam" in the December, 1983, issue of *QST*.

The Landstorfer antenna does make a nice exercise in wave theory. Maybe somebody will come up with a more compact structure that will provide increased gain and be useful for construction on the HF bands. A step in this direction is the log-Yagi as outlined by Leo D. Johnson, W3EB, in "Log Yagis Simplified" (*ham radio,* May, 1983). **Wayne W. Cooper, AG4R**
Miami Shores, Florida

a VHF noise bridge

Measure resistances and reactances to within 3 ohms at 146 MHz

By A.E. Popodi, OE2APM/AA3K, Moosstrasse 7, Salzburg 5020, Austria

Although there is no need to enumerate the many merits of the noise bridge, it is not used as often as one might expect. One of the reasons for its limited acceptance by the Amateur community may be that one needs some knowledge of the impedance-transforming properties of transmission lines as well as modest mathematical skills in order to take full advantage of its great measuring potential.

Several excellent articles about noise bridges operating in the frequency range of 3.5 to 30 MHz have been presented in this magazine.[1,2] However, to date, nothing has been said about a VHF noise bridge.

The usefulness of a noise bridge depends mainly on its accuracy. Throughout the HF spectrum it is difficult to maintain a 1 percent accuracy over the relatively large frequency range of 3.5 to 30 MHz. In the VHF version bandwidth is small but the operating frequency is considerably higher, which poses the main difficulty. Conventional noise bridge construction with a variable resistor and a variable capacitor does not lead to success.

At VHF frequencies, a variable resistor of 200 ohms, for example, exhibits too much inductance. In addition, the sheer physical size of the variable capacitor in conjunction with the usually awkward position of its terminals make it practically impossible to provide interconnections without adding prohibitively large inductances, thus degrading performance severely.

This article describes the construction of a noise bridge for the 2-meter band (144 to 148 MHz) that has sufficient accuracy to permit credible and reproducible measurements on multi-element antennas or antenna systems consisting of antenna and transmission line. A practical example of determining the impedance of an antenna via its feedline from the shack is presented.

The four items that control noise bridge accuracy are the wideband transformer, the variable resistor, the variable capacitor, and the physical layout (wiring). Obviously, single-point grounding is mandatory. If we replace the variable resistor with a PIN diode and the variable capacitor with five parallel connected tuning diodes, it is possible to shrink the size of the noise bridge to a minimum and achieve almost perfect wiring. A certain weakness exists because of the rather poor RF properties of the SO-239 jack; because most Amateurs use the PL-259 plug, it was retained.

the wideband transformer

The transformer must have electrical symmetry and close magnetic coupling between primary and secondary windings. It consists of four tightly twisted No. 24 enameled copper wires with 0.5 mm diameter. Two opposite wires of the bundle are the primary and secondary windings. The bundle is threaded about 3.5 times through a toroidal core — for example, an Amidon T 50-10. The beginning and end of each winding must be marked with short pieces of sleeving material and then connected per **fig. 1**. It is important to have equal wire lengths between transformer terminals A and D and circuitry as well as between terminals B and C and point M. The toroidal transformer must be mechanically secured to the chassis with nylon hardware to prevent a change in calibration later on.

noise bridge diagram

The complete schematic of the noise bridge, shown in **fig. 2** consists of a separate noise source and bridge.

The noise signal is fed to the bridge via two equally long, thin coaxial cables 26.8 inches (68 cm) in length, corresponding to a half-wavelength at 146 MHz. The noise source is a 6.2-volt Zener diode whose noise spectrum is amplified and fed to transformer T2, which is identical to the bridge transformer T1. The secondary center tap of T2 is connected only to the two shields of the coaxial cables and not to the ground system of the noise generator. This results in reduced chassis currents and guarantees a very sharp null indication by the bridge. The noise generator is housed in a shielded box away

from the bridge section. The adjustable noise level results in improved null indication in some receivers and helps in finding the null. The center tap of the secondary winding of T1 is connected through a short piece of coax to the BNC connector (output to receiver). This BNC connector should be insulated from the chassis. The 22-μH choke provides the ground return for the diode current and the tuning voltage. A gear train is recommended for the R and C dials.

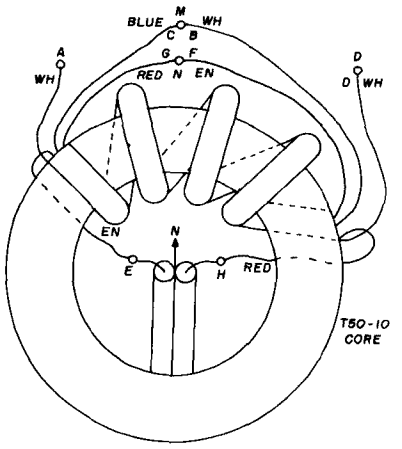


fig. 1. Winding interconnection diagram.

mechanical layout

Figure 3 shows the layout of the noise bridge components. A base plate of silver-plated brass serves as the common mounting surface for all bridge parts. Point A is a short, insulated standoff; points K and L are good high-frequency feedthrough capacitors. Additional bypass capacitors are provided in parallel with the feedthrough capacitors. All five diodes are mounted between two small metal plates that have five holes each. This diode package is then soldered between their respective points. Point M, the center tap of the secondary winding, is a plain feedthrough insulator. Point N is a heavy but short grounding standoff. It is also the grounding point for the two coaxial cables that carry the noise signal. The reference capacitor C_o must be soldered directly, with short leads, to the SO-239 jack.

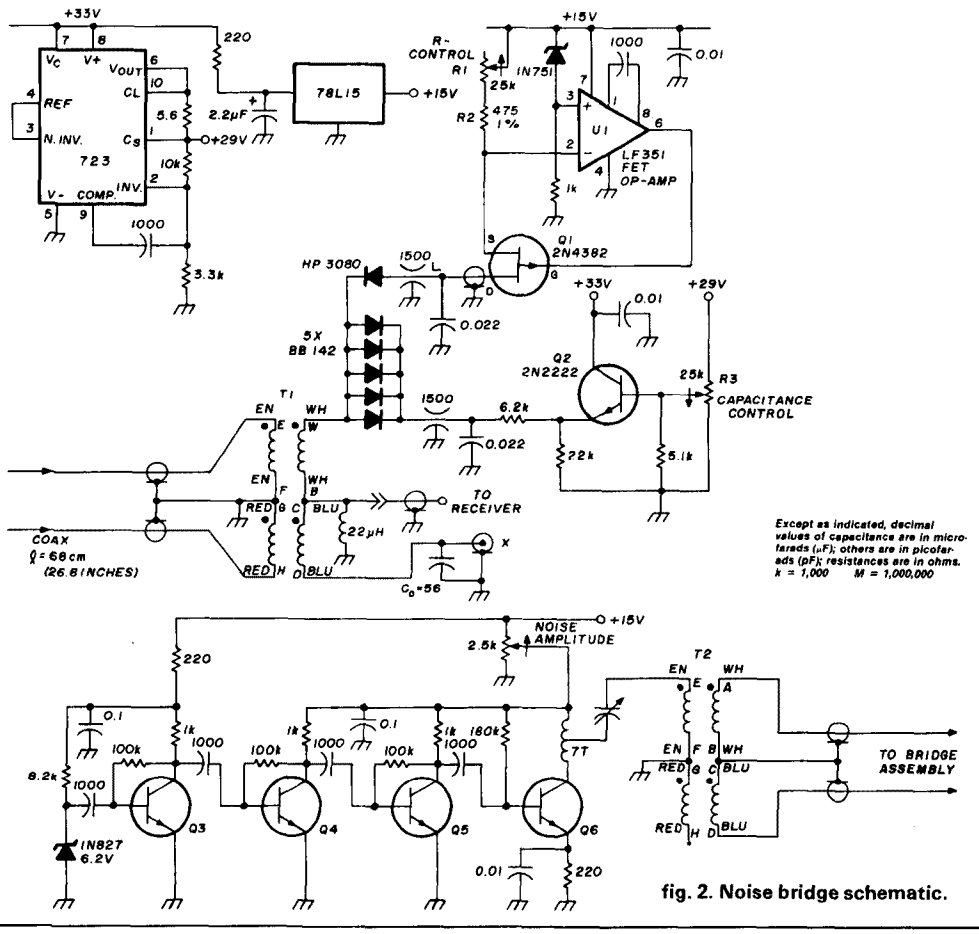


fig. 2. Noise bridge schematic.

control signals

The PIN diode requires a stable, constant current source. A P-channel FET, Q1, serving as constant current source, is controlled by the FET-input operational amplifier U1. At its input, a Zener reference voltage is compared with the voltage drop that the diode current generates across the variable resistor R1 and range-limiting resistor R2. The dial of resistor R1 is calibrated in ohms. Current range is about 0.2 to 10 mA, which corresponds to a diode resistance variation of 150 to 4 ohms. The connection to the diode should be made using shielded cable.

The five parallel connected tuning diodes require a variable DC voltage of 0.5 to 28 volts, controlled by potentiometer R3. The purpose of the 5.1-kilohm resistor is to linearize the (nonlinear) relationship between tuning voltage and capacitance. Transistor Q2 provides a low impedance and its base-emitter junction serves to compensate for the temperature coefficient of the tuning diodes. The 15- and 28-volt supplies must be regulated.

calibrating the resistance dial

The calibration of the R dial requires a little test jig to determine the relationship between diode AC resistance and DC current. This is shown in **fig. 4**. A signal generator supplies a 30-MHz signal to point A of an SPDT switch that may consist of a small metal bar that makes contact with either point B (non-inductive test resistor) or point C (PIN diode). The DC current can be delivered by the constant current supply described above or from an external DC source. A DVM is used to measure the current accurately. A second instrument is used to monitor the RF voltage at point A. The DC current must be adjusted until the voltages at B and C are equal. Three identical, non-inductive resistors of 50 ohms are required. The first measurement is made with 25 ohms (two resistors in parallel). The next readings are taken with 50, 100, and 150 ohms. This test jig must be built on a copper-clad board, with short leads to all critical points; a common grounding point is important. If the values of current are plotted on 3-by-3 decade logarithmic graph paper, the four points will lie on a straight line because of the logarithmic relationship between diode resistance and diode current. From this graph, the DC currents for all other resistance values can be read and the R dial can now be calibrated. (See **fig. 5**.)

calibrating the capacitance dial

This requires the measurement of the relationship between diode capacitance and DC voltage. The previously mentioned diode package can be measured with a capacitance bridge, but the applied voltage must not exceed the tuning voltage. Capacitance meters that use constant current charging cannot be used because of the 6.2-kilohm resistor.

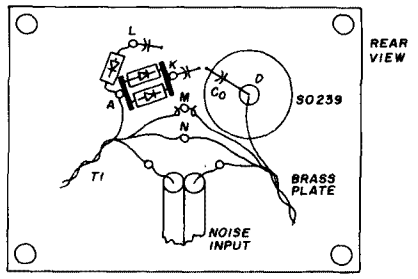


fig. 3. Bridge component layout.

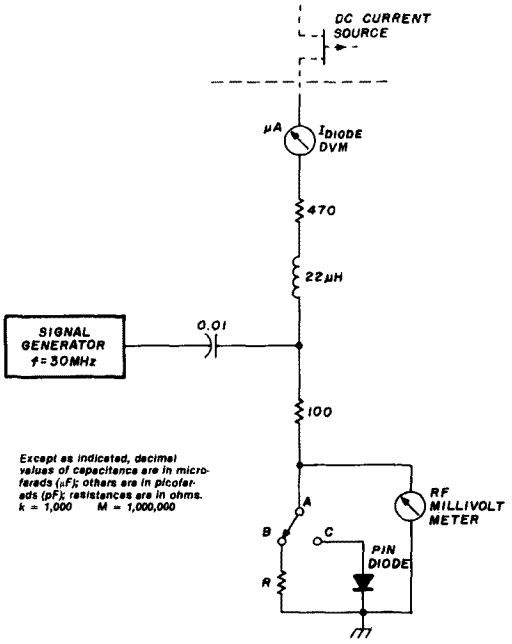


fig. 4. Test setup for resistance dial calibration.

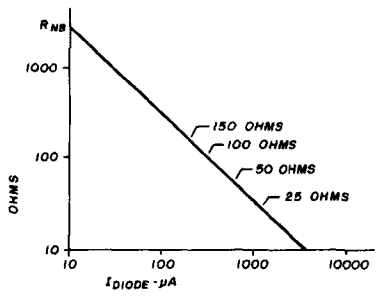


fig. 5. Resistance dial calibration chart.

Another method of measuring the capacitance in the noise bridge is to connect an inductor in parallel with the tuning diodes via a large DC blocking capacitor. The DC voltage may be set to 3.5 volts. The anode terminals of the diodes must be isolated from the noise bridge circuitry and a 100-kilohm resistor connected from this point to the ground to provide a DC return for the tuning voltage. With the inductor connected, measure the resonant frequency first. Then add a known capacitance, 33 pF, in parallel with the diodes and measure the new resonant frequency, repeating this with 68 pF. The values of diode capacitance and inductance can now be calculated. Ater removing these additional capacitors, calibrate the C dial. For example, a capacitance of 70 pF is related to a specific frequency that is selected by adjusting the tuning voltage. A recommended inductance is 0.26 μH, about 7 turns of diameter No. 18 AWG wire with 0.35-inch (9 mm) coil diameter.

calibrating the noise bridge

In order to understand the alignment procedure and interpret noise bridge readings properly, it is important to analyze the effect of parasitic inductances.

The equivalent circuit of a resistor that has self-inductance (as a result of excessive lead inductance) is shown in **fig. 6**. If we convert the series R-L circuit into its equivalent parallel R-L circuit, resistor R1 is always larger than R and L1 is larger than L. Since L1 is in parallel with the reference capacitor C_o, it reduces its apparent value and the noise bridge variable capacitor must be set to a smaller value than C_o in order to balance the bridge. Therefore, if we measure a resistor with self-inductance, the noise bridge indication is a higher R value and a capacitance value smaller than C_o. As an example, at 145 MHz, the noise bridge will measure a 50-ohm resistor at 51.34 ohms and with 3.5 pF less capacitance than C_o, if the series inductance is 9 nH. This happens to be the inductance of a 1-centimeter length of No. 28 wire.

Figure 7 shows the actual electrical circuit of the noise bridge X terminal area. L is the inductance between the center pin of the SO-239 connector and transformer, including leakage inductance. The reference capacitor C is assumed to be non-inductive. R is the load resistor. **Figure 8** shows the calculated resistive and capacitive components of the impedance measured between points A and B as a function of R, under the assumption that L=2.5 nH and C=58 pF. It is interesting to note that for resistance larger than 23 ohms the apparent capacitance is *larger* than C. This capacitance enhancement is caused by the inductance L, which is in series with C, creating a series L-C tuned cirucit that operates below its resonant frequency. We also see that a 50-ohm resistor appears to be only 38.8 ohms. If we reduce R from 50 to 25

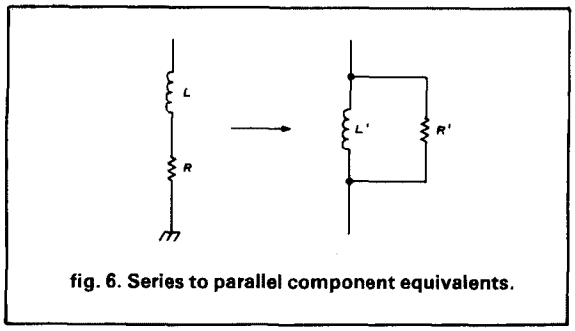


fig. 6. Series to parallel component equivalents.

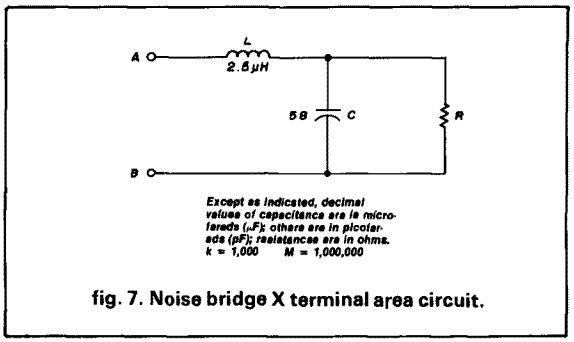


fig. 7. Noise bridge X terminal area circuit.

ohms, the resistance decreases fairly linearly, but the capacitance decreases non-linearly from 64.5 to 60.1 pF. In other words, the resistance variation results in a resistance *and* capacitance change.

From the above it may appear that building an accurate noise bridge is very difficult, if not impossible. However, if we provide the same inductance of 2.5 nH in the other branch of the noise bridge, the same capacitance enhancement occurs, but in the opposite direction and the above 38.8-ohm resistor becomes 50 ohms again. Therefore, the criteria for correct noise bridge operation are as follows:

1. A resistance change at the X terminal must not affect the capacitance reading.

2. A capacitance change at the X terminal must not affect the R reading.

3. A capacitance variation at the X terminal must be equal to the change of the variable capacitor. If we add, for example, a 33-pF capacitor in parallel with the reference capacitor and if the variable capacitor is increased by only 25 pF to obtain bridge null, then the capacitance enhancement at the X terminal side is too small. We must therefore add inductance to the X terminal side.

How well these criteria are met depends on transformer construction and alignment. Despite the low impedance level of the bridge circuitry, parasitic capacitances also affect noise bridge accuracy. PIN- and tuning diodes may also introduce self-inductance.

$$Z_L = R_L + jX_L = Z_o \frac{Z_o \sinh (a + jb) - (R_{IN} + j X_{IN}) \cosh (a + jb)}{(R_{IN} + j X_{IN}) \sinh (a + jb) - Z_o \cosh (a + jb)} \quad (1)$$

$$Z_{IN} = R_{IN} + j X_{IN} = Z_o \frac{(R_L + jX_L) \cosh (a + jb) + Z_o \sinh (a + jb)}{(R_L + j X_L) \sinh (a + jb) + Z_o \cosh (a + jb)} \quad (2)$$

a — Nepers b — Radians

$$R_L = Z_o \frac{Z_o R_{IN} (1 + A^2 + B^2) - A (Z_o^2 + R_{IN}^2 + X_{IN}^2)}{Z_o^2 + (A^2 + B^2)(R_{IN}^2 + X_{IN}^2) + 2Z_o (BX_{IN} - AR_{IN})} \quad (3a)$$

$$X_L = Z_o \frac{Z_o X_{IN} (1 - A^2 - B^2) + B (R_{IN}^2 + X_{IN}^2 - Z_o^2)}{Z_o^2 + (A^2 + B^2)(R_{IN}^2 + X_{IN}^2) + 2Z_o (BX_{IN} - AR_{IN})} \quad (3b)$$

$$R_{IN} = Z_o \frac{(R_L + Z_o A)(Z_o + AR_L - BX_L) + (R_L + AZ_o)(AX_L + BR_L)}{(Z_o + AR_L - BX_L)^2 + (AX_L + BR_L)^2} \quad (4a)$$

$$X_{IN} = Z_o \frac{(X_L + BZ_o)(Z_o + AR_L - BX_L) - (R_L + AZ_o)(AX_L + BR_L)}{(Z_o + AR_L - BX_L)^2 + (AX_L + BR_L)^2} \quad (4b)$$

$$D = 1 + e^{4a} + 2e^{2a} \cos (2b) \qquad A = \frac{e^{4a} - 1}{D} \qquad B = \frac{2e^{2a} \sin (2b)}{D}$$

a — Nepers b — Degrees $e = 2.71828$

Parallel to series conversion:

$$R_S = R_P \frac{X_P^2}{R_P^2 + X_P^2} \qquad X_S = X_P \frac{R_P^2}{R_P^2 + X_P^2} \quad (5)$$

Series to parallel conversion:

$$R_P = \frac{R_S^2 + X_S^2}{R_S} \qquad X_P = \frac{R_S^2 + X_S^2}{X_S}$$

Table 1. Conversion formulas.

recommended calibration procedure

All parts must be added directly in parallel to the reference capacitor C_o, using the shortest possible leads.

1. Connect a non-inductive, 1/4-watt resistor in parallel to C_o.

2. Set the resistance dial to 50 ohms and the capacitance dial to a value equal to the sum of C_o and the SO-239 socket capacitance (in our case, 59.5 pF). If there is no noise null, slight bending or reshaping of the transformer leads will help. A small piece of ferrite held close to one of the wires indicates where to correct.

3. Add a 33 pF capacitor to C_o. if the capacitance dial must be increased by less than 33 pF to obtain bridge null, additional inductance must be added at the X terminal side (this inductance may consist of a simple 0.157 inch, wire loop). If necessary, the taps on the center balancing inductors shown in **fig. 3**, may be changed. This balancing inductor consists of three turns of bare wire, its center point soldered to the feedthrough insulator. If equality cannot be achieved, the ratio of the two capacitance values (which is larger than unity) can be used as an instrument correction factor so that any reading on the capacitance dial is multiplied by that factor. This improves bridge accuracy.

4. Add another 50-ohm resistor in parallel to the first one. The R dial should indicate 25 ohms, and the C dial setting should not change.

5. Adding a 33-pF capacitor to the first 50-ohm resistor should not affect the R dial indication.

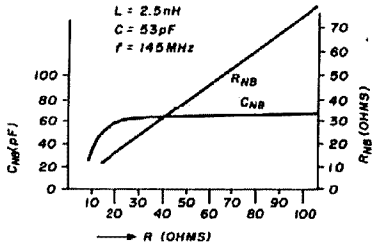


fig. 8. Calculated resistive and capacitive components of the impedance measured between points A and B.

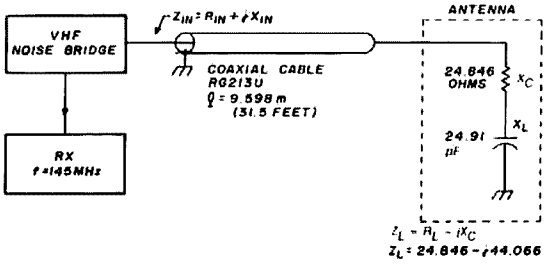


fig. 9. Noise bridge application.

measuring admittances with the noise bridge

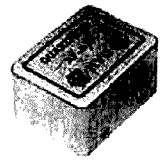
The ideal location for admittance measurements is on the C_o side of the SO-239 connector, but this is not convenient. Measurements on the other side of the connector are easier to make. However, the additional inductance of the 0.748-inch center pin reduces accuracy, especially if adapters must be used. For example, the difference in R and C readings is about 1.5 ohms and 2 pF if a 50-ohm BNC termination resistor is measured, using a PL to BNC adapter. However, the error is small.

measuring antenna impedance: example

The block diagram of the test setup is shown in fig. 9. It is assumed that the antenna may be represented by the series connection of a 24.846-ohm resistor and a 24.91-pF capacitor. The noise bridge is connected to the cable input with the 2-meter receiver serving as bridge null indicator.

At this point we must digress and remember that we can calculate the load impedance Z_l of a transmission line from the (measured) input impedance Z_{IN} if the cable parameters — attenuation, length, and characteristic impedance — are known. Likewise, we can determine Z_{IN} if Z_l is known. These conversions can be made with

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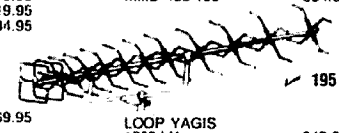
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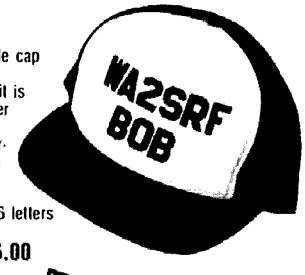
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Smith charts, but formulas are more accurate and convenient, especially if a programmable calculator or computer is available. **Table 1** is a listing of the required formulas. Formulas 1 and 2 can be used if you have a programmable calculator with complex operating mode (for example, an HP-15C). Formulas 3 and 4 lead to the same result, but are more cumbersome to use.

Since the noise bridge readings are in parallel resistance values and the formulas assume series configuration, we must convert parallel-into-series circuits and vice-versa. These formulas are also listed in **table 1.**

The most important cable parameter is length, represented by the term b in the formulas; it must be known fairly accurately. One method of determining the cable length, besides actually measuring it with a scale, is to disconnect the antenna from the coax and connect instead a suitable known resistor. You can then calculate with formulas 2 or 4 the cable input impedance Z_{IN} and compare it to the input impedance that was measured with the noise bridge. If there is a difference, try other values for b until agreement is reached.

In our example, assume the cable length is not precisely known and estimate a length of 31.3 feet (9.55 meters). (**Figure 9** shows the exact length as 31.49 feet, or 9.598 meters.) The cable length corresponding to one wavelength at 145 MHz is

$$\frac{3 \cdot 10^8}{145 \cdot 10^6} \cdot 0.66 = 1.3655 \text{ m or } 4.48 \text{ feet} \quad (1)$$

where the factor 0.66 is the velocity constant of the cable. Since we have assumed a length of 31.3 feet (9.55 meters), the value of b in formulas 1 and 2 becomes

$$b = 360° \frac{9.55}{1.3655} \cdot \frac{1}{57.296} = 43.943 \text{ radians} \quad (2)$$

The factor 57.296 converts degrees into radians. If we use formulas 3 and 4, b must be expressed in degrees.

The value of a, the attenuation of the 31.3-foot (9.55-meter) cable, can be found from a handbook. Assuming 3 dB per 100 feet (30.48 meters), we obtain:

$$a = \frac{3 \cdot 9.55}{30.48} \cdot 0.115 = 0.1081 \text{ Nepers} \quad (3)$$

The factor 0.115 converts dB into Nepers. The characteristic impedance Z_o is assumed to be 50 ohms.

Now terminate the coax with a known resistor. Select a value that renders readings well within the range of the noise bridge. Obviously, values too close to 50 ohms must be avoided. The resistor must be accurately measured with a DVM. Solder the non-inductive (preferably 1/4-watt) resistor, keeping leads as short as possible, between the center pin and flange of an SO-239 socket and then plug it into the coax. The load impedance in our example is:

$$Z_L = 101.3 + j0$$

(all values are expressed in ohms). Using formulas 2 or 4, the input impedance Z_{IN} is calculated to be with $a = 0.1081$ and $b = 43.943$ to:

$$Z_{IN} = 87.299 + j\,4.046$$

This is the input impedance measurement that should be indicated by the noise bridge.

The noise bridge readings (taken at 145 MHz) are: $R_{NB} = 77.5$ ohms and $C_{NB} = 65$ pF. Since the capacitance setting for a pure resistor is 59.5 pF, the capacitance difference is 5.5 pF. This represents, at 145 MHz, a capacitive reactance of $-$j 199.57 ohms. The noise bridge readings are therefore 77.5 ohms in parallel with $-$j 199.57 ohms. Converting into its series equivalent, using formula 5, we have $Z_{IN} = 67.34 - j26.15$. Comparing this with the above calculated value for Z_{IN} from Z_L for another set of parameters, a and b, we find, after a few trials, we that $a = 0.109$ and $b = 44.3406$ render the best agreement between measured and calculated Z_{IN}. This value is

$$Z_{IN} = 69.67 - j\,27$$

This compares favorably with the noise bridge measured input impedance of $Z_{IN} = 67.34 - j\,26.15$.

Having found the correct cable parameters, reconnect the antenna (in our example, the test load) and measure the input impedance Z_{IN} with the noise bridge. The readings are: $B_{NB} = 40$ ohms, $C_{NB} = 83.5$ pF. The actual capacitance is $83.5 - 59.5 = 24$ pF. Capacitive reactance is $-j45.73$. Converting into its series equivalent with formula 5, we obtain: $Z_{IN} = 22.66 - j\,19.82$. This is the measured cable input impedance with the antenna connected. This value must be inserted in formulas 1 or 3 to calculate the antenna impedance Z_L, using $a = 0.109$ and $b = 44.3406$. The result is: $Z_L = 27.195 - j\,43.97$. This is the measured antenna impedance.

As shown in **fig. 9**, the actual load impedance is $Z_L = 24.846 - j\,44.066$.

This is a remarkably good result, considering the high operating frequency and the fact that the assumed values for cable attenuation and characteristic impedance may not be correct. However, errors in a and Z_0 are much less significant than errors in b.

In this example, the actual load consists of a 103-ohm resistor in parallel with an 18.9-pF capacitor, which is the parallel equivalent. This permits a more inductance-free connection, thereby reducing the instrumentation error.

summary

The construction of a VHF noise bridge that is accurate enough to permit credible and reproducible measurements on antenna systems is possible if a PIN diode serves as variable resistor and tuning diodes replace the variable capacitor. The component layout must be such that no wiring interconnections are necessary. Undesirable ground currents that degrade the null indication are

avoided using single point grounding, a brass mounting surface, balanced noise injection, and proper transformer construction.

For antenna measurements via the feedline, we must know its exact length. If the length is not known, it can be found from measurements of the cable input impedance with the noise bridge and by comparing this value with the calculated input impedance, using a known termination resistor. One advantage of this method is in the fact that the calculated value of *b* represents the actual cable condition and includes its velocity constant, which does not have to be known. All necessary formulas to calculate load impedance from input impedance, and vice-versa, are presented as are network conversion formulas.

If the above method cannot be used because we cannot connect the test load, or if the cable length is not known accurately enough, the noise bridge readings are still the exact load impedance that the transceiver "sees." This load is the antenna impedance, which may be transformed by the transmission line into a completely different value, depending on cable length.

If you want to measure impedances directly at the X terminal, be careful to avoid undesirable inductances and capacitances. Series connection of several adapters will degrade accuracy. Generally, a resistor with series inductance (or self-inductance) measures higher; the capacitance reading is below the reference value.

To circumvent the deficiencies of the SO-239 connector, measure the unknown impedance using a coaxial cable half a wavelength long at the frequency of interest, with properly integrated PL-259 plugs on both ends. Thus, the SO-239 connector becomes part of the coax geometry and does not affect accuracy. Components must be connected with the shortest possible leads, preferably using an SO-239 jack that is then connected to the cable.

Possible sources of error are the instrument tolerance in the noise bridge and discrepancies in the assumed values for cable attenuation and characteristic impedance. The validity of the formulas may also be questioned. But according to numerous tests — and bearing in mind the statistical distribution of errors — the RMS error of this VHF noise bridge is less than 3 ohms in both the resistive and reactive components.

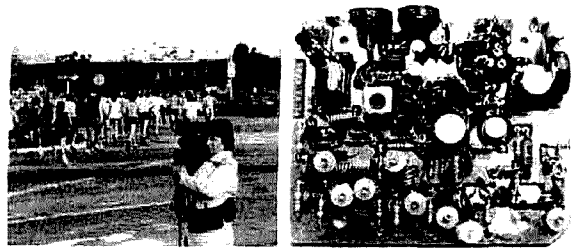
There is still room for improvement of the presented design, possibly in transformer construction, bridge alignment, and in establishing electrical symmetry between the X terminal and reference branch of the noise bridge.

references

1. R.A. Hubbs, W6BXI, and A.F. Doting, W6NKU, "Improvements to The RX Noise Bridge," *ham radio*, February, 1977, page 10.
2. Forrest Gehrke, K2BT, "A Precision Noise Bridge," *ham radio*, March, 1983, page 50.

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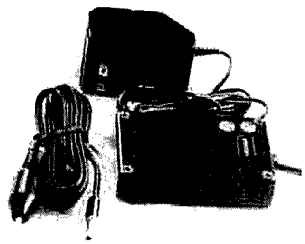
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artificial transmission lines

Construction of striplines and hybrid couplers using hand tools

Impedances can be matched in a variety of ways; some examples are the well known 1/4-wave stub and its lesser known cousin, the series-section matching line. For OSCAR work, where circular polarization is required, the branch line hybrid is ideal.

In most matching methods, transmission lines of non-standard impedances are required. The easy way out is to parallel two or more coax cables to come close to the required impedance; the harder way is to fabricate transmission line from copper pipe and brass tubing. Series-section matching uses two or more lengths of different, but standard value impedance coax. The problem with these, however, is splicing them together.

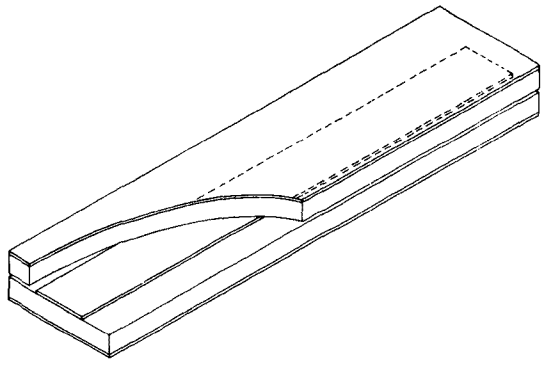


fig. 1A. Two printed circuit boards are assembled to form a stripline transmission line.

stripline

One form of transmission line used at UHF is called stripline. It is basically two printed circuit board strips. One PC board has a ground plane on one side and the transmission line center conductor etched on the other side. The other PC strip has a ground plane only. The two are assembled as in **fig. 1A**. To complete the self-shielding feature, such as in coaxial cable, a conductive foil covers the exposed edges of the dielectric. The impedance of a stripline transmission line impedance is directly proportional to the width, W, of the line and inversely proportional to the ground-plane spacing, b. The W/b ratio and the dielectric constant, ϵ, of the laminate determine the impedance of line (see **fig. 1B**).

But there's a problem with stripline. Printed circuit boards are fairly thin, and the center conductor width is small for the more popular line impedances. Thin traces are suitable for printed-circuit manufacturing but are rather awkward for the average ham to manufacture. The way to circumvent this problem is to widen the ground-plane spacing with Plexiglass® . The center line thus becomes proportionally larger for the same impedance (see **fig. 2**).

A quarter-wave artificial line was made with a printed circuit board exterior, two Plexiglas spacers and 0.002 inch (0.05 mm) brass shim stock for the center wire (**fig. 3**). Measurements were made with a General Radio admittance meter type 1602-A (**fig. 4**) at 150 MHz. The result for 0.25 inch, 0.125 inch and 0.062 inch Plexiglass is shown in **fig. 5**. Some advantages are obvious; for a line impedance of 50 ohms and using 0.25 inch (6.5 mm) spacers the center conductor width is 0.4 inch (10 mm), a dimension easily cut with average tools. But consider, too, the surface area. Most readers know about skin depth at radio frequencies. RF travels on the skin of a conductor. At

By Rudolf E. Six, KA8OBL, 30725 Tennessee, Roseville, Michigan 48066

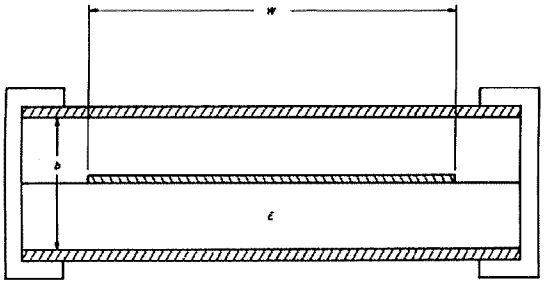


fig. 1B Impedance of the stripline is determined by width to ground-plane spacing, W/b and dielectric constant ϵ, of laminate.

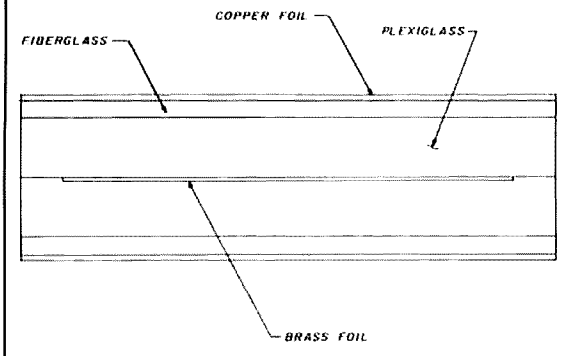


fig. 2. Ground-plane spacing may be widened with Plexiglass to avoid the problem of small center conductor width in common PC boards.

a depth, δ, ("skin depth") the current decreases to 37 percent of its surface value. For a copper conductor,

$$\delta = 2.6\sqrt{1/f}$$

where

δ = skin depth (inches)
f = frequency (Hertz).

As an example, at 150 MHz skin depth would be 0.00021 inch (0.005 mm). Thus, 0.002 inch (0.05 mm) shim stock is adequate for UHF. A width of 0.4 inch (10 mm) is equivalent to a round center conductor of approximately 0.25 inch (6.4 mm) diameter.

The resonant length depends on the dielectric of the Plexiglass:

$$V = 1/\sqrt{\epsilon}$$

where

V = velocity of propagation
ϵ = dielectric constant

and

$$L = \frac{3 \cdot 10^{10} \cdot V}{f}$$

where

L = electrical wavelength (cm)
f = frequency (Hertz)
V = velocity factor

The measured velocity factor was 0.60 or a dielectric constant of 2.78, close to the listed constant of 2.8 for Plexiglass. The Fiberglass part of the PC board is part of the line dielectric (ϵ = 4) and affects the total dielectric constant as the Plexiglass gets thinner. I used a velocity factor of 0.6 for the length calculation of quarter-wave stubs. The SWR, being the important factor, is flat over a wide frequency range. The hybrid coupler, however, has a sharp isolation bandwidth and several factors affecting resonant length, which are discussed later.

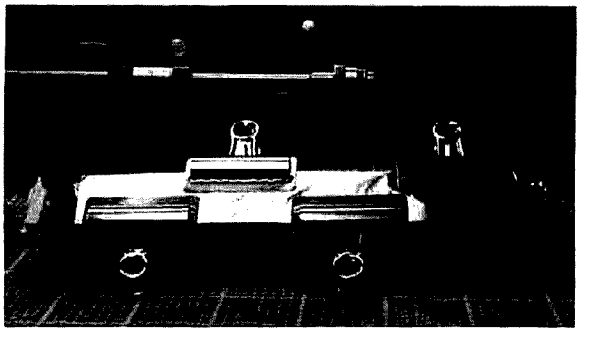


fig. 3. 1/4-wave stub fitted with a GR connector on the left and a 50-ohm termination made up of four film resistors on the right. Aluminum foil wrap stub are held together with paper clips.



fig. 4. Test setup using the GR 1602-A bridge.

construction of stripline

The center conductor is 0.002 inch (0.05 mm) brass. This is available as brass shim stock in sheets or rolls

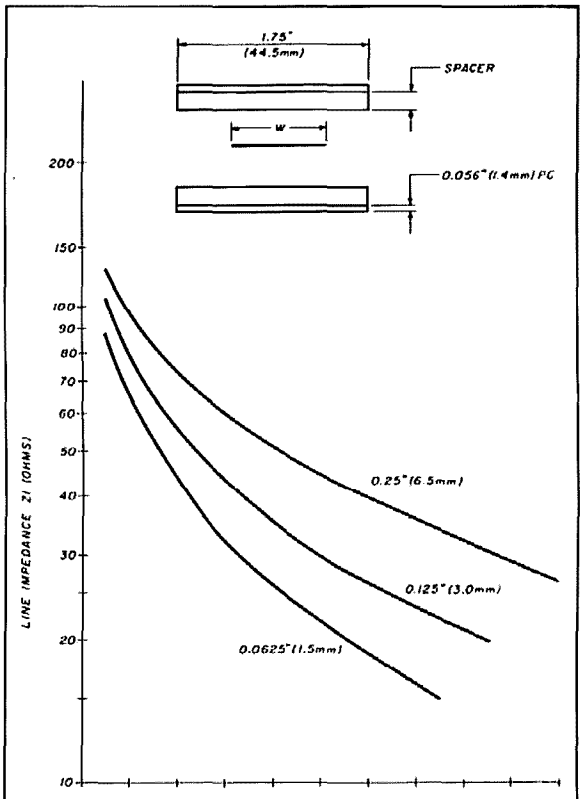


fig. 5. Test results for the stripline shown in fig. 3. Measurements were made with a GR 1602-A admittance meter at 150 MHz.

from machine tool supply firms. **Figure 6** shows the fixture used for accurately slicing the shim stock. The base is a thick slab of Plexiglass. Formica or bakelite would also work, but the material must be hard enough so an Xacto® blade won't cut into it. I bought two steel rulers and mounted one firmly to the Plexiglass. I attached the second ruler with two DeStaco hold-down clamps, also available from tool supply firms.

To cut an accurate strip, the shim stock edge is placed against the fixed ruler. The second ruler, with the shim underneath, is spaced to the required width with vernier calipers. The clamps are secured and an Xacto knife is used to slice the brass shim stock. It takes a little practice, but the results are much better than you'd get with scissors.

The plastic spacers are Plexiglass brand Lucite.® I used this well-known brand because of its widespread availability and the possibility of variations in the dielectric constant of other brands. Check the thickness of the Plexiglass; it does have some rather wide tolerances. The printed circuit board is standard 0.0625 inch (1.5 mm) G-10 glass epoxy. A table saw with a fine-tooth blade was used to cut the Plexiglass and the PC board.

To mount the RF connector, a hole is first drilled through one of the PC boards and both Plexiglas spacers. The bottom PC board is left undrilled. The diameter of the hole is the width of the mounting flange of the RF connector. I used BNC connectors because most of my test equipment uses BNCs. BNC connectors, however, are prone to be intermittent. Use UHF connectors for 2-meter work and N connectors for 70 cm and above. The TNC connector is excellent for low-power work at UHF. Not well known to Amateurs, the TNC is small and has low SWR characteristics.*

A small hole was drilled into the brass strip for soldering to the connector. This presented quite a problem since the spinning drill grabbed the thin brass, but the problem was solved by squeezing the brass strip between two pieces of scrap Plexiglass (**fig. 7**). The top piece has a hole predrilled to the right diameter. The brass strip is positioned accurately underneath, then the assembly is held firmly by hand and the hole is drilled through. The result is a nice clean hole in the right place.

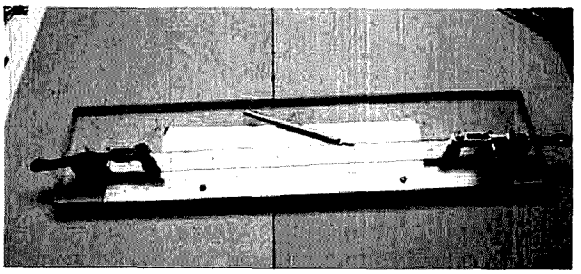


fig. 6. Tools for cutting brass shim stock.

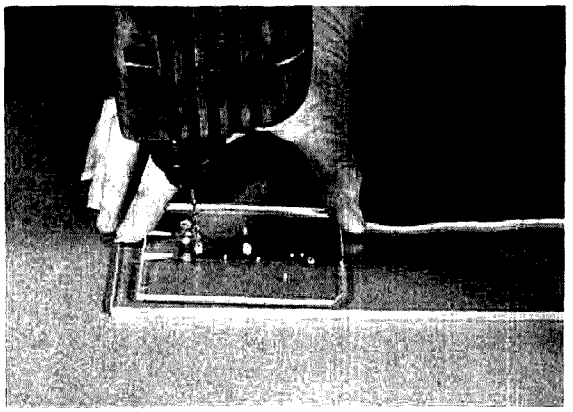


fig. 7. Drilling a hole in the brass shim stock. Two pieces of Plexiglass are used to guide the drill.

*Microwave Components, 11216 Cape Cod, Taylor, Michigan 48180.

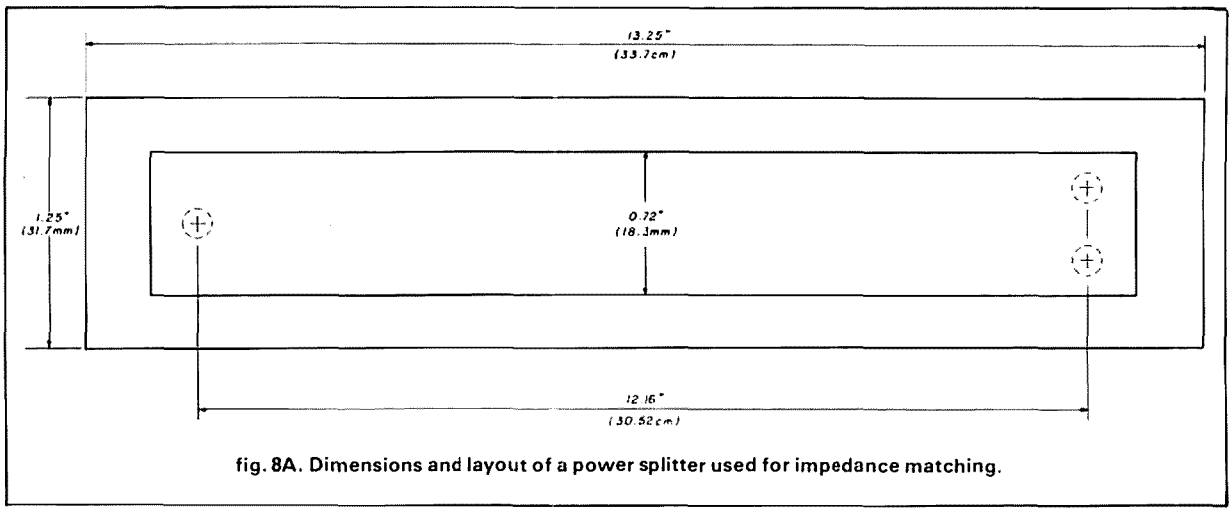


fig. 8A. Dimensions and layout of a power splitter used for impedance matching.

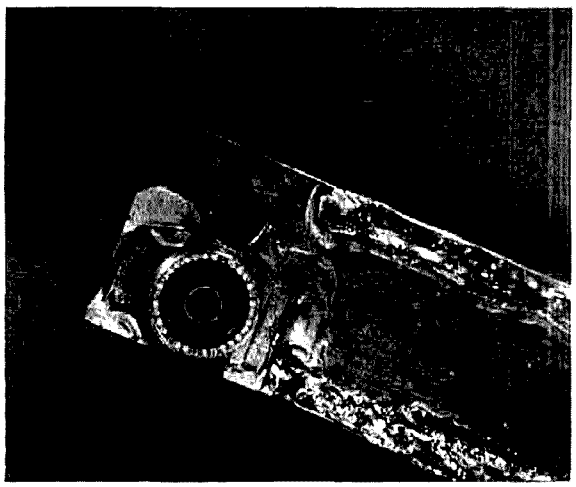


fig. 8B. UHF connector has been filed down to fit. Note solder access hole for the connector on the back side.

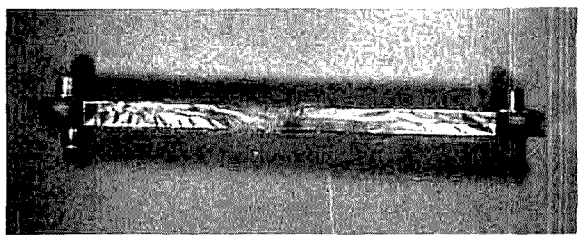


fig. 8C. A quarter-wave 2-meter power splitter.

The transmission line is ready for final assembly after the center conductor has been soldered to the RF connectors. I could have drilled a number of holes and held the transmission line together with nuts and bolts, but I found it easier to glue the whole assembly. I simply held everything together with C-clamps and wicked plastic glue along the edges with a hypodermic needle. The glue is methylene chloride.**

Avoid using plastic glue with acetic acid as an ingredient (vinegar odor); it is conductive. A variety of glues are available from commercial firms selling plastics. The outer conductor is completed by soldering 0.001 inch (0.03 mm) brass shim stock between both PC boards.

How wide should the line be made? Or in other words, how far away should the center conductor edge be from the outer shield? If the transmission line is too wide, it tends to be inductive; if it's too narrow, the line impedance is affected by the outer shield being

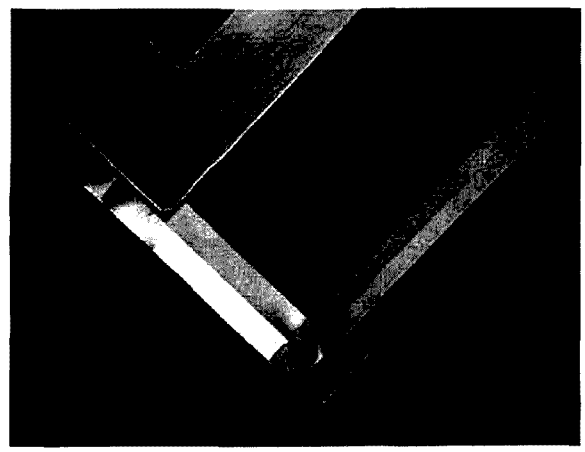


fig. 9. The pieces making up a hybrid coupler. When the strip was cut, a line was drawn with a needle to mark the center of the strip. This line was used for drilling the solder hole.

**Weldon 3", available from Industrial Polychemical Service, P.O. Box 379, Gardena, California 90247.

too close to the center conductor. I found the best distance between center strip and outer shield to be equal to the thickness of the spacers. Thus if 0.125 inch plastic spacers are used, the total width of the stripline is 0.375 inch (including the 0.0625 inch PC boards) larger than the width of the center conductor.

matching lines

A quarter-wavelength matching line is used to demonstrate the first practical use of these striplines. The impedance transforming properties of quarter-wave transmission lines are well known. Most Amateurs use them to couple antennas with an impedance different from that of the feed coax. Basically the input impedance, Z_a, of a quarter-wave line terminated in a resistive impedance, Z_b, is

$$Z_a = Z1^2/Z_b$$

where Z1 = impedance of quarter-wave line, rearranging

$$Z1 = \sqrt{Z_a Z_b}$$

This means that the matching line, commonly called

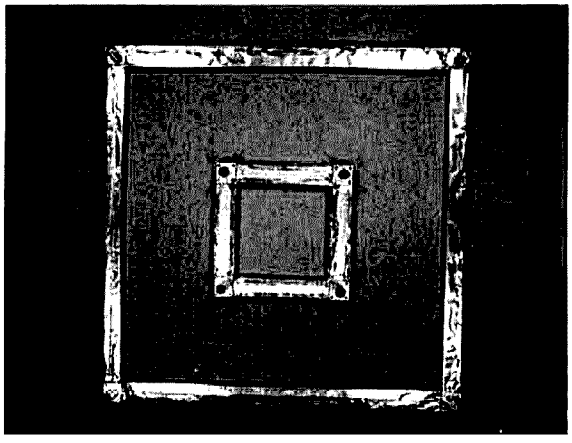


fig. 10. Completed hybrid couplers for 2 meters and 70 cm.

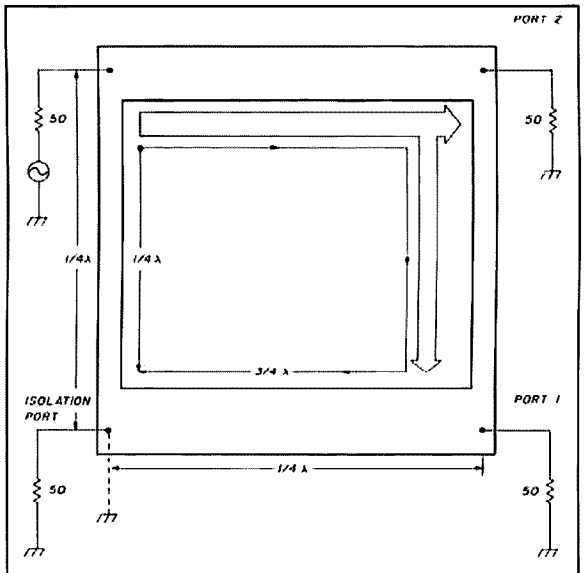


fig. 12. Isolation port is necessary to keep unwanted currents out of the system, resulting in isolation between ports.

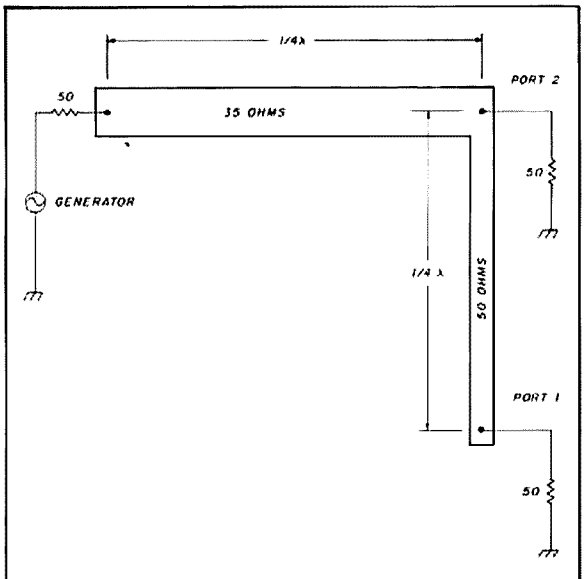


fig. 11. Impedance and phase shift relationships in a hybrid coupler.

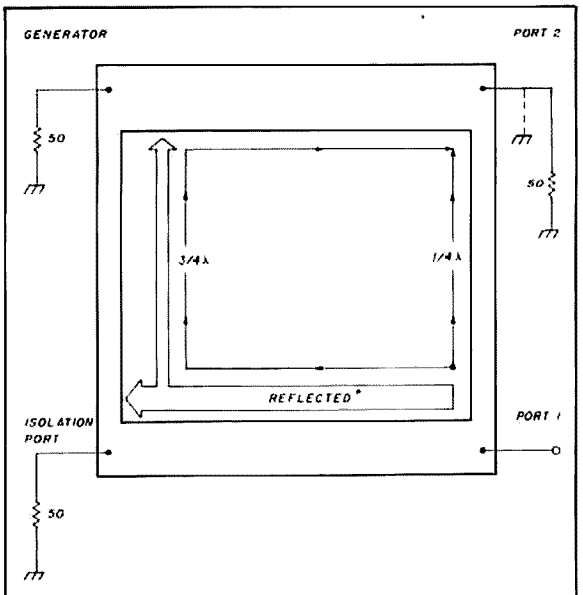


fig. 13. Reflected current due to poor port matching.

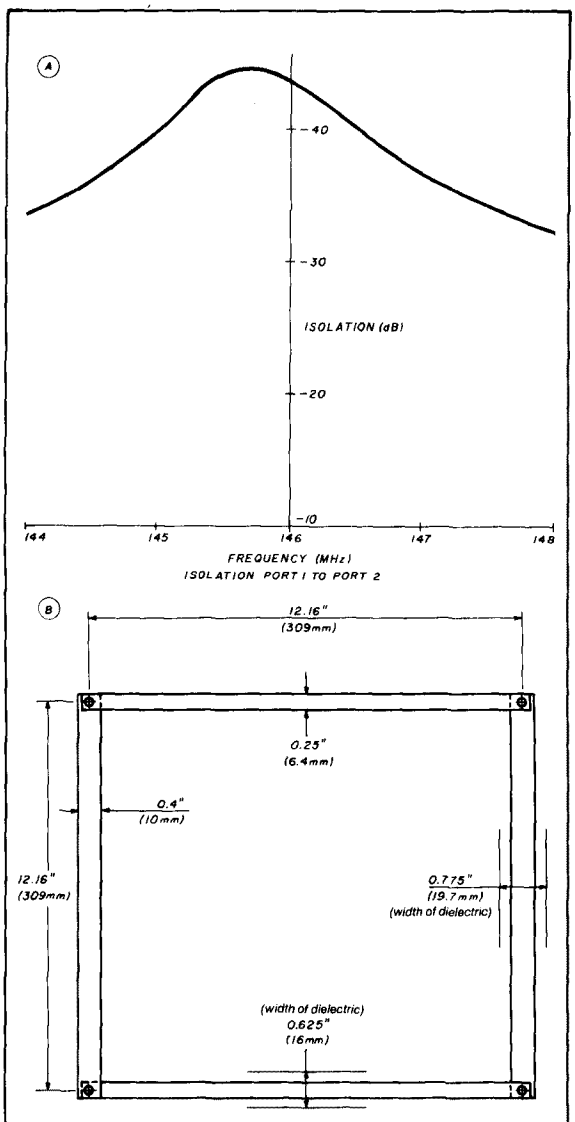


fig. 14. 2-meter hybrid coupler: (A) Port 1-2 isolation; (B) dimensions.

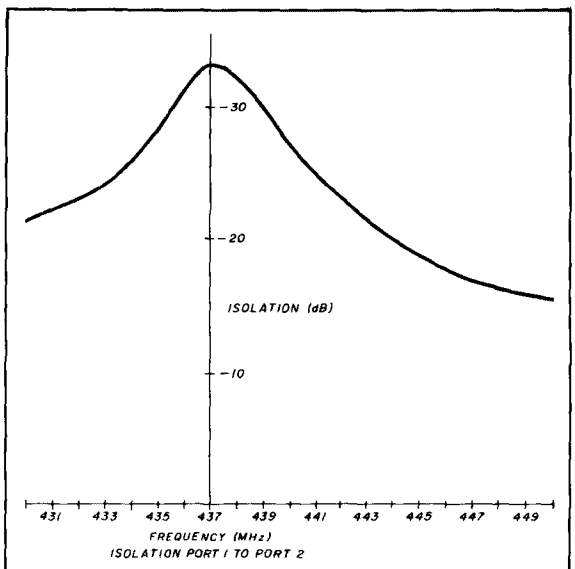


fig. 15. Test results of the 2-meter hybrid coupler operating in the 70-cm band.

a Q section, must have an impedance, $Z1$ to match an antenna Z_a to coax Z_b. If two antennas of 50 ohms are paralleled, the impedance is 25 ohms, and a Q section of 35 ohms is required to match this combination to 50-ohm coax.

The dimensions and layout for such a power splitter are shown in **fig. 8A**. With 0.25 inch (6.4 mm) spacers, the SWR was less than 1.1:1 from 144-148 MHz. A UHF connector was soldered to each side of the line to make it possible to solder them side by side. Even so, their mounting base had to be filed down to make them fit **fig. 8B**. The mounting hole for the UHF connector was drilled through both PC and Plexiglass strips so the center conductor could be soldered to each connector from the back. A completed power splitter is shown in **fig. 8C**. Various types of power dividers are covered in **reference 1.**

hybrid couplers

Made up of interconnected transmission lines, the branch-arm hybrid coupler provides a 3-dB power split and a 90-degree phase shift between both outputs (**fig. 9**). Complete units for 2 meters and 70 cm are shown in **fig. 10**. This is ideal for OSCAR 10 communications, in which two Yagis are mounted at right angles to each other and fed out of phase by 90 degrees. Depending on which antenna is fed first, we can get right-hand or left-hand circular polarization. The hybrid coupler can be used in reverse to combine the signal returning from the satellite. Ernie Franke, WA2EWT, wrote an excellent article on this device.[2]

I constructed a hybrid-coupler from coax for the receiving end of my OSCAR 10 station. Three coax cables, including the outer shields, had to be soldered to each RF connector. Because the physical result didn't exactly excite me, I tried to come up with a better way of making these couplers; this led to the work presented here. (Although the basic idea behind these couplers is covered in this article, I would also recommend that anyone interested read parts 1 and 2 of reference 2.)

The quarter wavelength 50-ohm line parallels port 1 and port 2 with a total resistance of 25 ohms at port 2 (**fig. 11**). Note that port 1 has a 90-degree phase shift from port 2. The 35-ohm line is used to convert the

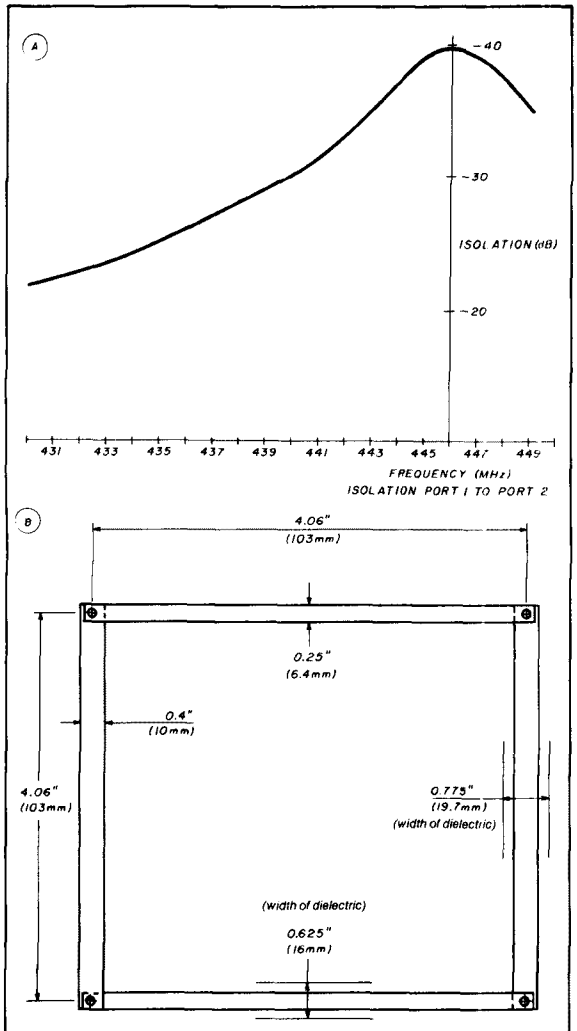


fig. 16. 70-cm hybrid coupler: (A) Port 1-2 isolation; (B) dimensions.

25 ohms back to 50 ohms. Adding a second leg (**fig. 12**) sets up the following conditions. The current from the generator divides itself between port 2 and port 1 as before. A much smaller current also flows towards the isolation port from the generator following two paths. The long way around is one-half wavelength longer, or 180 degrees out of phase. These signals arrive in opposite polarity and cancel. This virtual short at the isolation port sets up two shorted quarter-wave stubs, one toward the generator and one toward port 1. Their high impedance prevents current from flowing into the isolation port.

Now what happens when reflected current flows back into port 1 because of poor port matching? The reflected current is absorbed by the isolation port and the generator. Again a small current flowing toward port 2 by way of two paths creates a virtual short at

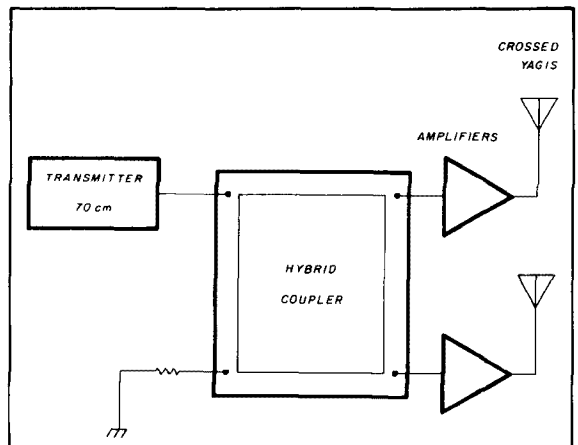


fig. 17. General arrangement of author's 70-cm OSCAR 10 station.

port 2. The resulting shorted quarter wave stubs prevent current flow from port 1 to port 2. The result is isolation between ports (**fig. 13**).

The first hybrid coupler constructed was for 2 meters. With care in fabrication, excellent isolation can be achieved between ports (**fig. 14A**). Inside and outside cuts on the Plexiglass and PC frame were made with a table saw. Make sure the result is square. The distance between the RF connectors was marked accurately before a pilot hole was drilled. Good isolation depends on equal distances between ports. Even if the frame isn't perfectly square, make sure these distances are equal. Note the width of the frame legs, 0.625 inch (16 mm) and 0.775 inch (19.7 mm). One might think that a fractional value would be hard to duplicate; buy a pair of inexpensive plastic vernier calipers and practice on a piece of scrap before making the final cut. All the hybrid couplers used 0.125 inch (3 mm) Plexiglass. The completed assembly was held together with eight small C-clamps. Methylene Chloride was wicked between the pieces with a hypodermic needle (a small brush or Q-tip® would also work).

After approximately an hour's drying time, the shield is completed by soldering on the 0.001 inch (0.03 mm) brass shim stock.

Width of the stripline also affects the frequency resonance of the coupler. The line becomes inductive with increasing width. Since 0.125 inch (3 mm) Plexiglass is used, the line width is 0.375 inch (9.5 mm) wider (including the 0.0625 inch [1.5 mm] PC boards) than the brass center conductor. All these factors are much more apparent in the hybrid coupler than in the simpler power splitter because the hybrid coupler tunes more sharply.

I also ran some tests with the 2-meter hybrid coupler at the third harmonic in the 70 cm band. Each leg

then becomes 1 and 1/2 wavelengths. The frequency tripled almost perfectly (**fig. 15**) and again with good isolation.

The second coupler was made for the 70 cm band. Dimensions are shown in **fig. 16**. As the coupler becomes smaller, care in construction becomes even more important. The center of the frame had to be cut with a sabre saw because the unit is too small for a table saw blade. A file was used to finish the dimensions. This unit will be used for the transmitting end of OSCAR 10.

Presently the rig has one low-power amplifier and the output drives two crossed Yagis with a matching arrangement. I can now double the output by adding an amplifier. Impedances are 50 ohms, and the advantage here is that I can tune antennas for minimum SWR with a standard 50-ohm SWR bridge.

conclusions

The aim of this article has been to present a simple method of constructing matching devices of high quality, in which connector problems commonly encoun-

Appendix A (See fig. 5.)
The information below is provided for the benefit of readers who wish to make their own graph for the data shown in fig. 5.

	Zt termination (ohms)	W spacers inch (mm)	1/Zc conductance (μmhos)		susceptance (μmhos)		Z1 impedance (ohms)	
0.25 inch	50.2	0.047(1.19)	3.4		0		121.5	
(6.5 mm)	50.0	0.097 (2.46)	5.1		0.4		99.0	
	50.3	0.150 (03.8)	7.2		0.15		83.6	
	50.1	0.203 (5.16)	9.5		0.3		72.6	
	50.1	0.245 (6.22)	11.4		0.3		66.3	
	50.1	0.297 (7.54)	13.9		0.5		60.0	
	50.2	0.348 (8.84)	16.35		0.9		55.4	
	50.2	0.403 (10.2)	19.2	20.8	1.4	.6	51.1	51.1
	50.1	0.445 (11.3)	22.2	17.5	1.4	1.6	47.6	46.9
	50.2	0.498 (12.6)	25.2	16.25	2	2.5	44.6	45.1
	50.2	0.550 (13.9)		14		1.3		41.9
	50.2	0.600 (15.2)		12.55		1.5		39.7
	50.2	0.650 (16.5)		11.15		1.3		37.4
	50.2	0.750 (19.0)		9.25		1.3		34.1
	50.2	0.850 (21.6)		7.55		1.3		30.8
	50.2	1.000 (25.4)		6.00		1.3		27.4
0.125 inch								
(3.0 mm)	50.3	0.047 (1.19)	4.6		0.5		104.6	
	50.2	0.097 (2.46)	7.8		0.6		80.2	
	50.2	0.150 (3.81)	11.55		0.8		65.9	
	50.2	0.203 (5.16)	16.1		0.6		55.8	
	50.3	0.245 (6.22)	20.3		1.0		49.8	
	50.1	0.297 (7.54)		15.1		0.7		43.5
	50.1	0.348 (8.84)		12.4		0.6		39.4
	50.1	0.403 (10.2)		10.2		0.9		35.7
	50.2	0.445 (11.3)		8.8		0.9		33.2
	50.2	0.498 (12.6)		7.6		1.0		30.9
	50.2	0.550 (13.9)		6.55		1.0		28.7
	50.2	0.650 (16.5)		5		1.0		25
	50.2	0.750 (19.0)		4		1.2		22.4
	50.2	0.850 (21.6)		3.25		1.4		20.2
0.0625 inch								
(1.5 mm)	50.1	0.047 (1.19)	6.4		0		88.5	
	50.1	0.097 (2.46)	11.85		0.2		65.1	
	50.1	0.150 (3.81)	19.15		0.2		51.1	
	50.1	0.203 (5.16)		15.0	0			43.4
	50.1	0.245 (6.22)		11.7		0.1		38.3
	50.1	0.297 (7.54)		8.9		0.4		33.4
	50.1	0.403 (10.2)		5.5		0.5		26.2
	50.1	0.498 (12.6)		4.0		0.6		22.4
	50.1	0.600 (15.2)		2.9		0.9		19.4
	50.1	0.750 (19.0)		1.9		0.9		15.4

1/4-wave measuring above 50 ohms

$Z1 = \sqrt{Zt Zc}$

1/2-wave measuring below 50 ohms

$Z1 = \sqrt{2.5\ Zt/Zc}$

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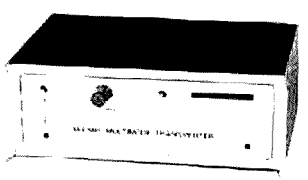
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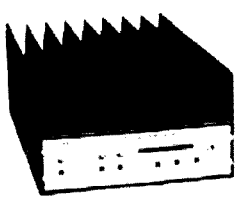
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Appendix B. Data for figs. 14, 15, and 16.

2-meter hybrid coupler Port 1-2 isolation.

143.95 MHz	– 34 dB
144.39	– 36
144.86	– 39
145.13	– 41
145.88	– 44
146.37	– 41
146.8	– 38
147.34	– 35
148.04	– 32

2-meter hybrid coupler (operated in 70-cm band) Port 1-2 isolation.

430.065	– 20 dB
431.695	– 23
433.475	– 25
434.42	– 27
435.55	– 30
437.5	– 33
438.95	– 30
440	– 27
440.97	– 25
442.64	– 23
444.035	– 20
447.035	– 18
449.82	– 15

70-cm hybrid coupler Port 1-2 isolation.

449 MHz	– 35 dB
448	– 37
447	– 39
446	– 40
445	– 39
444	– 37
443	– 35
442	– 33
441	– 31.5
440	– 30
439	– 29
438	– 28.5
437	27.5
436	26.5
435	25.5
434	25.5
433	24
432	23.5
431	23
430	22.5

tered in making matching systems are eliminated.

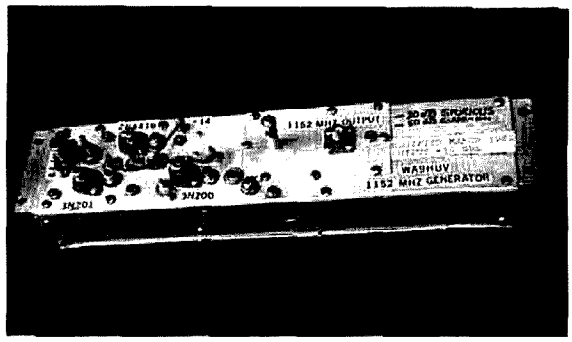
Moving up to the GHz ranges could be accomplished by using styrofoam rather than Plexiglass in a ring hybrid; the size problem could be avoided by operating on an odd harmonic such as shown above.

references

1. George Chaney, W5JTL, "Extended/Expanded Power Dividers," *ham radio*, October, 1984, page 73.
2. Ernie Franke, WA2EWT, "The Branch-Line Hybrid," *ham radio*, Parts 1 and 2, April (page 107) and May (page 93), 1984.

ham radio

New type of PLL beats the phase noise barrier



1968 model 1152 MHz signal source.

low-noise phase-locked UHF VCO

part 1: the noise problem

Much has been written about voltage-controlled oscillators. In a recent *ham radio* article Hans Roensch, W0DTV, described a free-running UHF VCO, correctly acknowledging that such an oscillator cannot be used effectively in narrowband systems where low phase noise is required.[1] W0DTV's article referenced an earlier article of mine, which included a description of a way to phase lock oscillators to reduce noise.[2]

One of the major disadvantages of the phase-locked loop I described in reference 2 is the lack of AFC, requiring that the VCO be set on frequency manually to effect phase lock. Under adverse conditions, drift of the oscillator's tuned circuit may exceed the hold-in capability of the loop, causing the VCO to unlock or jump to another phase-lock point. In this scheme the oscillator's frequency is divided by a factor of 40 to make it compatible with the phase detector. This, as we shall see, is another disadvantage, which adds 32 dB of excess noise to the VCO.

To make this a more practical system, two additional features are needed: automatic frequency control (AFC) and UHF phase detection to avoid the need for prescaling.

background

A free-running VCO intended for use as a receiver's local oscillator (LO) in the 1215-1300 MHz band will generally exhibit a very broadband phase noise characteristic. Without the benefit of crystal control, its spectrum may be so wide that the carrier is not readily identifiable on a spectrum analyzer because the deviation factor is so large. While the degree of phase modulation is relatively minor on the HF and VHF bands, as frequency is increased the modulation angle becomes proportionally larger. Self-excited TV oscillators, for example, often sound like buzz saws on a UHF receiver as they wander around. Something better than this is obviously needed for use as a local oscillator in a narrowband receiver intended for reception of NBFM, SSB, or CW in the 1215-1300 MHz band.

The 1296-MHz transceiver I put together fifteen or twenty years ago was built around a 1152-MHz crystal-controlled signal source. I added a 144-MHz SSB signal to the 1152-MHz generator in an upconverter to develop the 1296-MHz SSB exciter signal. Conversely, I used the 1152-MHz generator as the LO in a downconverter and a 2-meter receiver as the IF. This is a scheme in general use by many UHFers. Note that twice the 1152-MHz oscillator frequency is 2304 MHz, a fact of considerable interest to those operating both bands.

The original 1296-MHz exciter employed a crystal-oscillator-multiplier chain driving a step-recovery di-

By **Norman J. Foot, WA9HUV**, 293 East Madison Avenue, Elmhurst, Illinois 60126

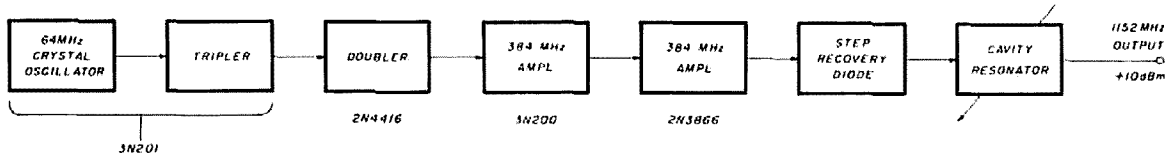


fig. 1. Block diagram of 1968-model 1152 MHz crystal oscillator multiplier chain.

ode (SRD) multiplier associated with a cavity resonator to develop the 1152-MHz local-oscillator signal. A 64-MHz crystal oscillator was followed by a tripler, a doubler, and two stages of 384-MHz amplification. The second amplifier, a 2N3866, was used to drive the SRD. The power output from the cavity resonator was +10 dBm. Most of the circuitry was relatively straightforward, but the SRD and cavity resonator circuits were quite complicated mechanically. Many contacts were made on 1296 MHz using this signal source for the exciter and downconverter, even before the 1296-MHz band was as well populated as it is today. **Figure 1** is a block diagram showing the crystal-oscillator multiplier system.

instrumentation

Since that time I have acquired an HP-8551/851B spectrum analyzer, which gives me the opportunity to look more closely at the oscillator's noise and spurious performance. The frequency spectrum showed spurious signals about 30 dB below the carrier and amplitude and phase noise at or below the noise floor of the spectrum analyzer. An oscillogram of its spectrum is shown in **fig. 2**, with a photo of the unit.

The 8551 can measure 70 dB below the carrier in a 1-kHz bandwidth. Phase and amplitude noise of an oscillator can be observed on this spectrum analyzer if the noise is great enough; however, for phase noise measurements of a clean and highly stable source, the spectrum analyzer leaves much to be desired. Nevertheless, it does represent a benchmark against which the noise and spurious signal characteristics of other oscillators can be compared.

Why be concerned with oscillator spectral purity? Because sideband amplitude and phase noise can convert directly into the receiver's passband and limit overall sensitivity under multiple signal conditions.[3,4] For example, when a weak signal is being detected in the presence of a stronger signal, oscillator phase noise — beating with the strong signal — appears at the mixer's output and enters the IF even though the strong signal may not. If the phase noise of the local oscillator is great enough, the desired signal may be completely buried in noise. Therefore, it is important that the phase noise of the local oscillator be made as small as possible. Low phase noise is one of the most important characteristics of an oscillator.

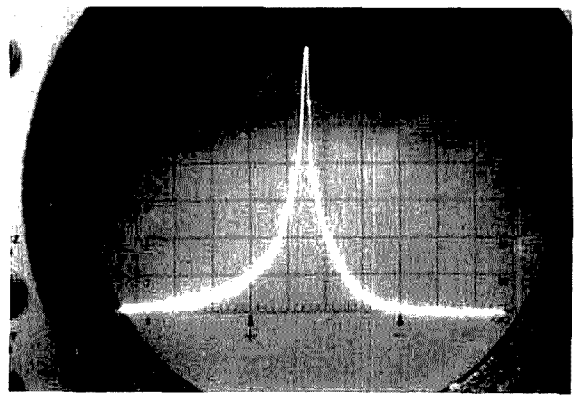


fig. 2. Oscillogram of spectrum of crystal oscillator multiplier chain. BW = 1 kHz. Horizontal scale 30 kHz/cm; Vertical scale 10 dB/cm.

specifications

The VCO described here was intended to replace the 1152-MHz crystal-controlled oscillator-multiplier chain discussed above. The ground rules I set were to achieve performance as close to that of the original unit as possible and to accomplish this in a simple manner. The oscillator would be operated directly at 1152 MHz without the need for frequency multipliers or cavity resonators and would be phase locked to a crystal reference signal in a relatively simple phase-lock loop. AFC would be included to avoid the possibility of any frequency ambiguity.

phase-locked loops

The theoretical aspects of phase-locked loops have been well covered in the references listed at the end of this article. It might be a good idea to do some reviewing before proceeding further. In particular I suggest reading (or re-reading) the two articles by Craig Corsetto, WA6OAA.[5,6] For the most part, I will use Corsetto's notation to make this article easy to follow.

phase noise at VHF

It is less difficult to design a PLL around a digital phase detector than an analog one, especially when the VCO operates in the HF or VHF bands. But going beyond VHF into the UHF region introduces problems

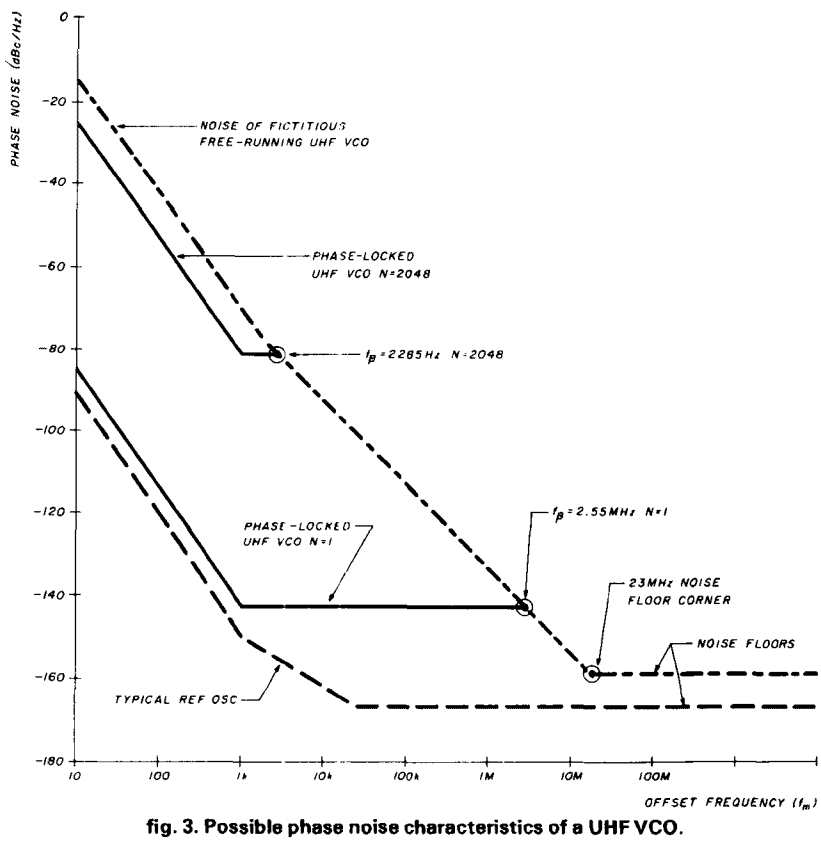


fig. 3. Possible phase noise characteristics of a UHF VCO.

that call for a considerable amount of ingenuity to keep phase noise low. The alternative would be to use multiple loops, but this increases the circuit complexity considerably.

To start out, I used a CD4046 phase/frequency detector; this required that the frequency of the VCO be prescaled (divided) by a large number so it would be compatible with the CD4046's upper frequency limit. I used an RCA CA3179 prescaler followed by a 74LS191 divider. The prescaler operated in the divide-by-256 (UHF) mode, and the 74LS191 was programmed to divide by eight. Total division was 2048, providing 562.5 kHz to the phase detector. By itself, this circuit was effective in providing AFC and phase detection, but, as we shall see, the noise introduced due to the division factor would be unacceptably high.

loop bandwidth

The loop bandwidth, $f\beta$, is a key factor in terms of phase noise because frequencies less than $f\beta$ are under the control of the PLL while those greater than $f\beta$ are not. $f\beta$ is the frequency corresponding to unity open-loop gain:

$$f_\beta = \frac{K_\phi K_{vto}}{2\pi N}$$

K_ϕ = *phase detector sensitivity in volts/radian.*

K_{vto} = *42 MHz/volt, voltage tuning sensitivity.*

N = *2048, oscillator division factor.**

For the *CD4046*, $K_\theta = 0.7$. Plugging in these values yields $f\beta = 2285$. The significance of this number will now be discussed.

It is possible to calculate to a fairly good approximation the phase noise of a free-running oscillator as a function of sideband frequency, using the following relationships:[7,8,9]

$$\pounds(f_m) = 10 \log \left\{ \frac{FKT}{2P_{av}} \left[1 + \left(\frac{f_0}{2f_m Q_L} \right)^2 \right] \right\} dBc/Hz$$

This equation is not as formidable as it may seem. The first term is Johnson noise divided by the oscillator's power output, which is 1/SNR (1/signal-to-noise ratio). F is the noise factor of the active device, K is Boltman's constant, and T is the absolute temperature, ^oK.

The second term represents the manner in which noise is distributed on each side of the carrier frequen-

*I have departed from Corsetto's notion by defining N in the usual manner as the divisor, a number equal to or greater than 1. This should help avoid confusion.

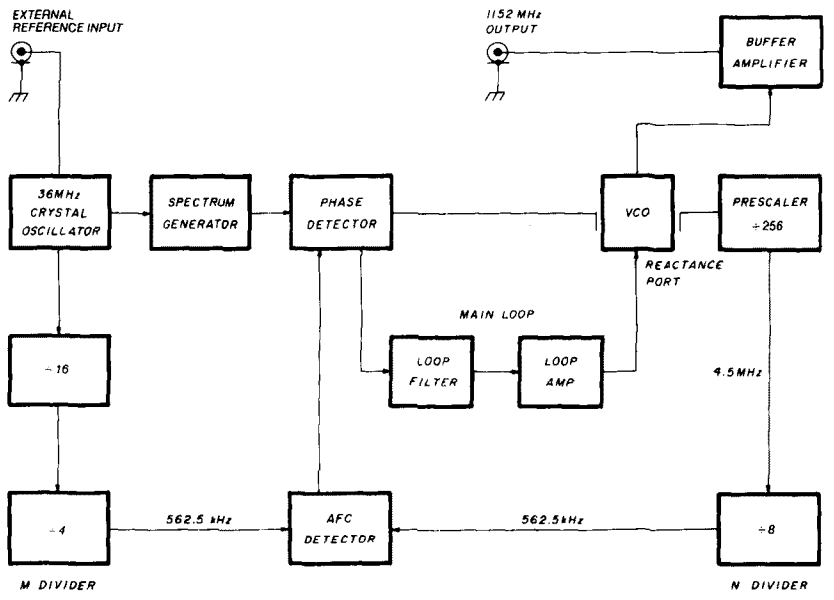


fig. 4. Functional block diagram, new 1152 MHz generator.

cy. The term f_m is the distance away from the carrier that a measure of phase noise might be made. Note that when f_m is very large, the second term approaches unity, and

$$\pounds\ (f_m)\ approaches\ 10\ log\left[\frac{1}{2\ SNR}\right]$$

This is generally referred to as the single-sideband noise floor.

Going away from the carrier, VCO noise decreases until it intersects the noise floor at $f_m = f_o/2Q_L$. At a VCO frequency f_o = 1152 MHz and loaded Q, Q_L = 25, the noise floor corner is 23 MHz. (As a rule of thumb, $Q_L = 0.2\ Q_U$, where Q_U is the unloaded Q.)

closing the loop

Using the various values of f_m, the dotted curve of **fig. 3** was drawn, including the noise floor, noise corner, and unity gain band edge, $f\beta$. When the loop is closed, the noise departs from the 20-dB/octave slope and tends to become constant from $f\beta$ to the carrier. The solid curve of **fig. 3** represents the noise response of a fictitious VCO phase locked to a reference oscillator. Because N is so large, VCO noise near the carrier is relatively high.

The importance of $f\beta$ cannot be overemphasized. As the loop bandwidth is widened, more and more phase noise comes under the influence of the loop, making for a quieter VCO.

One way to widen the loop bandwidth is to reduce N. Unfortunately, the upper frequency limit of the CD4046 is about 2.5 MHz, based on a propagation time of 200 ns between terminals 15 and 13. The lower limit on N is therefore about 450, giving

$$f\beta \cong 10\ kHz$$

This reduces in-band phase noise by about 15 dB, which is better, but still not good enough.

phase noise characteristics

The phase-noise floor of a well designed crystal oscillator is about 165 dB below the carrier at offsets beyond 20 kHz but increases rapidly at lower frequencies as shown in **fig. 3.**[10] This curve was extrapolated from various manufacturer's specifications and includes low-frequency flicker noise. Since the noise of a VCO locked to this reference is increased by $20\ log\ N$ above the noise of the reference, the best that can be expected of the VCO under these conditions is about −80 dBm/Hz at offsets below $f\beta$. Notice that the phase-locked noise level is only 10 to 15 dB lower than for the free-running mode, because N is so large and $f\beta$ so low. The curve of the free-running VCO was calculated and has been verified only indirectly. While the curve is referred to as "fictitious," it may in fact be quite representative of the VCO's actual performance. Use of such a VCO as the LO in a communications receiver would surely degrade overall performance, especially in a multi-signal environment. An alternative would have to be found.

phase/frequency detector — part of the solution

The CD4046 is an interesting device because it is a

combination phase and frequency detector. In this dual role, it first steers the VCO toward the desired frequency (i.e., provides AFC) and then phase locks it to the reference. While the 4046 is not the only means to this end, it is far simpler than most other schemes. However, because it requires the use of a prescaler, phase noise is relatively high at UHF. What would happen if *somehow* N could be made equal to unity? $f\beta$ would then become 2.55 MHz, providing a theoretical reduction in phase noise of over 60 dB! This would put the VCO noise in the vicinity of −150 dBc/Hz and allow the noise specification of the UHF oscillator to be met.

In an earlier *ham radio* article I wrote about a programmable HF receiver that uses direct synthesis ($N=1$).[11] For AFC I added a counter, digital comparator, and a DAC (digital-to-analog converter). This receiver is still the workhorse of my station. It occurred to me recently that I might be able to use a similar direct phase detection scheme to provide phase locking and low noise at UHF, together with a CD4046 for AFC purposes. The question was, could these two circuits be made to work together? If so, this would finally solve the noise problem.

the final scheme

The UHF PLL that finally evolved is illustrated in functional block form in **fig. 4.** The main loop includes a phase detector operating directly at 1152 MHz. The auxiliary loop includes the CD4046, which provides AFC. A "picket fence" of reference signals[12] 36 MHz apart in the vicinity of 1152 MHz, is fed from the spectrum generator to one port of the phase detector. Another port receives a sample of output RF from the oscillator. The third port receives a frequency-steering signal from the CD4046. The output of the phase detector feeds a loop filter and amplifier driving the reactance port of the VCO.

With this configuration, the noise of the VCO is basically the same as that of the reference oscillator. Assuming that the noise curve of the reference shown in fig. 3 is accurate, then the VCO phase noise should be −90 dBc/Hz near the carrier, dropping down to −150 dBc/Hz 10 kHz from the carrier and beyond. These are very respectable numbers indeed.

frequency programming

Using the numbers associated with **fig.** 4 establishes 1152 MHz as the UHF oscillator's only operating frequency. However, a number of other options have been made available:

- Provisions have been made to enable selection of $N = 7$, 8, or 9, corresponding to VCO frequencies of 1008, 1152, or 1296 respectively ($f_{VCO} = 144N$).
- A VXO can be used instead of the fixed-frequency crystal oscillator to vary the UHF oscillator's frequency by a small amount, typically ±1.0 MHz.
- Other output frequencies may be selected by changing the crystal frequency. Given a desired UHF output frequency, the corresponding crystal frequency will be as follows:

$$f_x = \frac{64\, f_{VCO}}{2048} \qquad for\ N = 8$$

or $f_x = 0.03125\, f_{VCO}$

For example, if the VCO frequency is to be 1116 MHz, $f_x = 0.03125 \bullet 1116 = 34.875$ MHz.

As another example, suppose the UHF oscillator is to be used as the LO for an OSCAR 10 upconverter. If the SSB generator is at 28 MHz, $f_{VCO} = 463\ MHz$ and $f_x = 14.46875\ MHz$. Of course, in this situation both the crystal oscillator and UHF oscillator inductances would have to be increased appropriately.

If you are planning operation on the new 902-928 MHz band, and if you use 29 MHz as the IF, a high-side VCO will operate at 933 MHz and $f_x = 29.15625\ MHz$, or a low-side VCO will operate at 875 MHz with $f_x = 27.34375\ MHz$.

The circuits described here are by no means all-inclusive. The basic circuitry can easily be applied to other frequencies and other schemes, including fully-synthesized systems. For example, a frequency synthesizer could be substituted for the reference oscillator; however, this is beyond the scope of this article. The objective here is based on the concept that most Amateur UHF applications can be satisfied by providing a dedicated output frequency per individual circuit board.

Next month: construction, testing, and performance.

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ham radio

PRACTICALLY SPEAKING ... Joe Carr K4IPV

using the multimeter

In last month's column we discussed various forms of multimeters Amateur Radio and other electronic hobbyists use in making electronic measurements. This month we'll discuss how these instruments are used in practical applications. The proper use of volt, current and ohmmeters — as well as how to make high voltage and RF voltmeter probes — will also be discussed.

how to connect meters

It's essential that you learn the correct way to connect meters into a circuit, because using these instruments incorrectly can lead to catastrophe. **Figure 1** shows the correct methods for connecting the various forms of a basic meter. Note well that voltage, current, and resistance aren't measured in the same ways; this fact leads to a high potential for damage to multimeters because the different functions are switch-selected. It's all too easy to "misconnect" a multimeter by switching ranges without first changing the position of the probes. The connections for voltmeters and current meters are shown in **fig. 1A.**

There are two simple rules to memorize:

1. Voltmeters are connected in parallel with the load.

2. Current meters are connected in series with the load.

Don't *ever* violate these rules! The second rule — always connect the ammeter in series and *never* in parallel — is especially important. If an ammeter is connected across the load, its low internal resistance will draw large current from the power supply of the circuit under test, and that current is usually much larger than the full-scale range of the meter. On an analog meter (a "pointer type" rather than digital) the pointer will often bend around the peg and a puff of smoke will waft from the edge of the case. The problem may be reduced in some digital multimeters (DMM), but is still present. Fortunately, most manufacturers now place fuses in series with their multimeter probes.

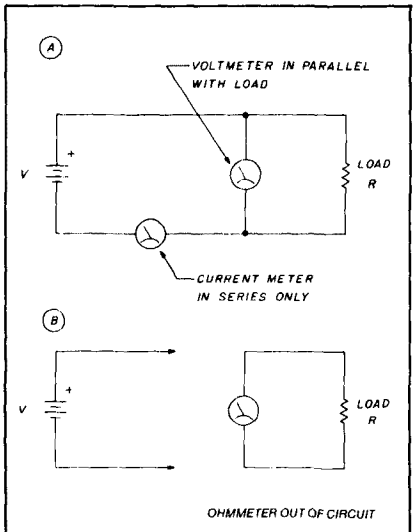


fig. 1. Correct methods for connecting meters to circuit.

The rule for ohmmeters is to disconnect the resistance being tested from the circuit even though the power is turned off. There are two reasons for this procedure: first, there may be parallel alternate paths for current to follow, and these will cause an erroneous lower reading; second, there may be current stored in capacitors in the circuit, and that current can be large enough to destroy the meters.

Power supply filter capacitors are particularly dangerous to meters. Although even professional electronics workers may ignore this rule, it's a good habit to follow even if disconnecting components is a bit of a nuisance.

voltmeter "errors"

I can recall an incident in a laboratory class in which an electrical engineering instructor couldn't tell a certain student why the voltages read in an experiment were considerably lower than called for in the lab manual — and lower than the results of other students. The reason for the error turned out to be loading of the circuit by the voltmeter.

A voltmeter has an input impedance. For a volt-ohm-milliammeter (VOM) type of instrument, the input impedance can be determined by the sensitivity rating in ohms per volt. Most good meters have a sensitivity of 20,000 ohms per volt, and some especially fine meters are rated at 100,000 ohms per volt. Many inexpensive imported meters, on the other hand, have a 1000 ohms/volt sensitivity, which is very bad. Incidentally, the sensitivity reflects the full-scale "natural" range of the meter movement used in the VOM. The 20,000 ohms/volt instrument uses a 50 μA movement, while the 1000 ohms/volt meter uses a 1 mA meter movement.

Figure 2 shows a sample circuit to illustrate the problem of loading. The circuit consists of a 10-volt source, V, and two series resistors, R1 and R2. What we need to know is the voltage across resistor R2 — that is, voltage V2. The correct voltage will be:

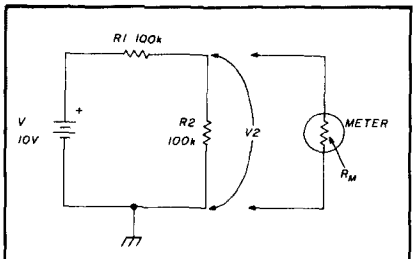


fig. 2. Simple circuit illustrates problem of loading.

$$V2 = \frac{V \cdot R2}{R1 + R2} = \frac{10\ volts \times 10K}{(100K + 10K)}$$

$$V2 = 0.909\ volts$$

Now, consider what happens when we connect the voltmeter across resistor R2. The total resistance of this branch of the circuit is now the parallel combination of R2 and the meter's input resistance. Consider a 1.5 volt full-scale setting for the meter. At this setting, the 20,000 ohm/volt model has an input impedance of 30,000 ohms, and the 1000 ohms/volt has an input impedance of 1500 ohms. When connected in parallel with R2 to measure the voltage drop across the resistance, the combination of R2 and the input impedance forces new values of "R2" of 7500 and 1300 ohms, respectively. These heavily loaded resistances reduce the measured voltages from 0.909 to 0.697 and 0.128 volts, respectively. These errors, which are substantial, clearly illustrate the reason for using a voltmeter with a high internal impedance.

high voltage probes

Most high voltage meters are really ordinary voltmeters with a multiplier resistor or voltage divider network added. **Figures 3** and **4** show two alternate methods for adding high voltage ranges to the standard multimeter. The HV probe of **fig. 3** is used on those meters that have a specified input impedance of 10 Megohms (for example, VTVMs). When the series resistance inside the probe is 990 Megohms, the voltmeter will read 1/100 the actual voltage. Thus, a 30,000 volt potential will read 300 volts on the meter. This type of probe is widely available from electronic supply houses, especially those that cater to television service technicians, who often use such probes in conjunction with regular voltmeters for measuring anode potentials on color TV sets.

The basis for the probe shown in **fig. 3** is the resistor voltage divider circuit, in which one element is the series resistance and the other is the voltmeter input impedance. These probes require an input impedance of 10 Megohms. However, modern electronic voltmeters, including FETVMs and DMMs, have input impedances much higher than 10 Megohms. For these cases we use a circuit such as the one shown in **fig. 4**. Again, when you buy an HV probe for your instrument it's likely to have this circuit inside; in fact, some of the "universal" probes on the market are little more than circuits such as the one shown in **fig. 4** with 100:1 or 1000:1 reduction ratios.

You can, if you prefer, make such a probe yourself. I once made one for a hospital electronics lab that was based on the old-fashioned model shown in **fig. 3**. We had the old probe, which mated with a traditional VTVM, but it didn't work on the new DMM with its advertised 1000-Megohm input impedance. Since the meter had a pair of banana jacks spaced 0.75-inch apart (a standard value), I mounted resistor R2 on a dual banana plug and soldered the combination to the ends of the old HV probe. The value of R2 was 1 Megohm, so with the 990 Megohms of the probe resistor I had a new probe with a 1000:1 reduction. We needed the probe for measuring the 12-15 KV anode potential of some elderly Sanborn scopes. In that case, a reading of 1.2 to 1.5 volts indicated the correct vlaue. In other applications, one could use the same 10 Megohms for R2 as was common for VTVM input impedances to make a 100:1 reduction HV probe.

A high voltage probe should be equipped with a good alligator clip for this connection. If you use an ordinary alligator probe, or a poor quality alligator clip for the common lead, then all bets are off — you could easily zap

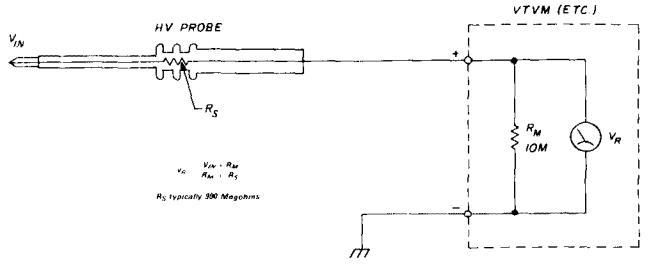


fig. 3. High-voltage probe is used on meters with specified input impedence of 10 Megohms.

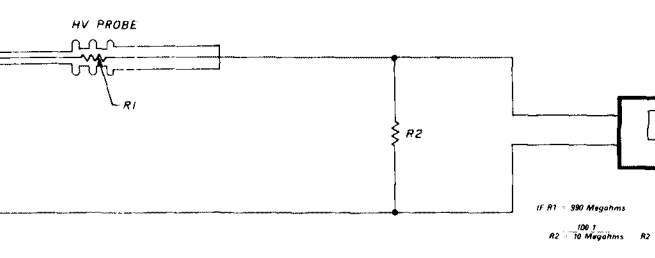


fig. 4. High voltage probes such as FETVMs and DMMs are likely to contain this circuit.

the meter. Damage to the meter is especially likely when it and the circuit under test are both grounded. If the common lead comes loose, the current from the high voltage circuit will try to find ground through the instrument.

meters in RF circuits

Very few multimeters are designed to operate in RF circuits. There are two problems: first, we sometimes need to measure DC voltages in the presence of large RF voltages; second, we might want at least a relative indication of the value of the RF voltage present in the circuits. Fortunately, we can build our own probes that serve both functions.

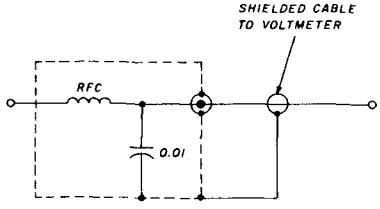


fig. 5. RF blocking probe allows measurement of DC voltages in circuits where RF is likely to be present.

The RF blocking probe shown in **fig. 5** is designed to allow measurement of DC voltages in circuits where RF is likely to be present, such as in a radio transmitter. The circuit inside the probe is a low-pass filter consisting of a 2.5 mH RF choke and a 0.01 μF capacitor. This circuit will block enough RF in most cases to permit a reading of the DC component. If there's still a problem, try two or three sections of the RFC/capacitor circuit to provide additional attenuation.

There are problems to be aware of when using the RF blocking probe shown in **fig. 5**, however. The most obvious is that the RFC and the capacitor might interact with LC elements in the circuit and could thus distort the readings. It's also possible to damage certain circuits by detuning them — as in detuning a plate tank circuit while the circuit is live, for example.

Another problem is that RF chokes have a resonant frequency. This frequency is a result of the inductance and interwinding capacitance of the choke. When using the probe at the choke's natural resonant frequency, it will act like any other resonant circuit and perhaps destroy itself.

RF voltmeter probe circuits are shown in **fig. 6**. Both of these are sometimes called "demodulator probes" because they'll demodulate AM signals. A service instrument called a signal tracer is little more than an audio amplifier with a low capacitance probe up-front to pick up audio signals. If we replace the low capacitance probe with one of the circuits shown in **fig. 6**, we'll be able to troubleshoot AM receiver RF and IF circuits. With the diodes shown, we can measure RF potentials up to about 50 or 60 volts peak. For greater potentials, use two or more diodes in series. The diodes, by the way, are old-fashioned germanium diodes. Although largely supplanted in the market by silicon diodes, there are still many 1N60 diodes used as video detectors in TV receivers. You can still buy 1N60 diodes under the ECG-109 and certain other "universal" replacement part numbers.

The probe shown in **fig. 6A** was used in a number of low-cost demodulator probes associated with signal tracers a few years ago. The circuit shown in **fig. 6B**, on the other hand, has been popular among Radio Amateurs for decades because a similar circuit was published in the ARRL's *Handbook*. Finally, there's another circuit, shown in **(fig. 6C)** that's also very popular, and can be calibrated in terms of RMS voltage.

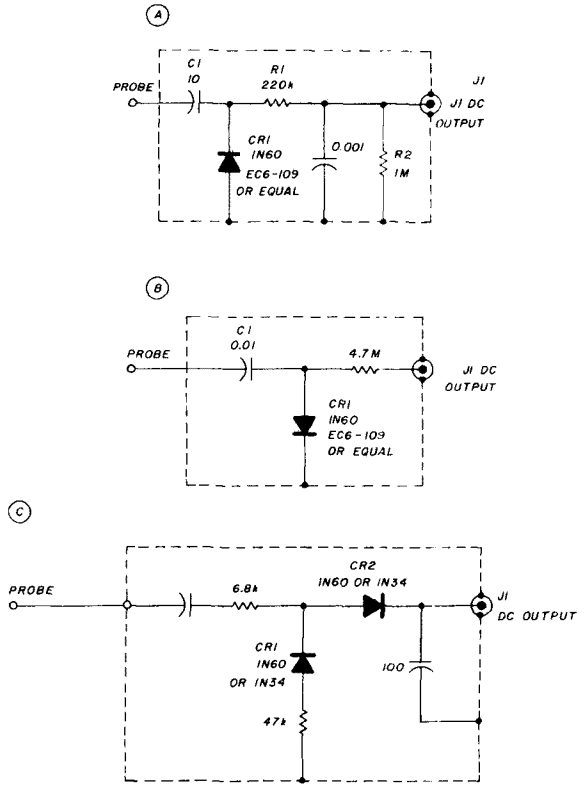


fig. 6. RF voltmeter probe circuits demodulate AM signals. (A) Low-cost demodulator was found in signal tracers; (B) Long popular with Radio Amateurs, this circuit was originally published in the ARRL *Handbook*; (C) Another popular circuit can be calibrated in terms of RMS voltage.

ham radio



KX0O with completed "Colorado Kilowatt" 6-meter amplifier. *Photographs courtesy W0MXY*

Easy-to-build 3CX1000A7 unit provides legal limit output

6-meter kilowatt amplifier

Several years ago I became interested in VHF contesting and soon realized that contests require the ultimate in VHF operating skill, knowledge of propagation, and the best equipment an operator can muster.

Six meters, located in the "gray zone" between HF and VHF, but exhibiting characteristics of both, has made or broken many a VHF contest effort. If it's open, working 6 meters requires topnotch skills to maintain rate and sift through pileups on or near 50.110. If 6 is dead, then high power, good antennas, and scatter techniques are the name of the game.

It was the urge to work the 6-meter scatter circuits — and the rigors of contesting — that made me consider obtaining a full legal-limit kilowatt for the band. The amplifier had to be able to deliver full power output without strain for hours. It had to be a trouble-free design; after all, who likes to doctor an amp during a band opening? Because of second harmonic problems in the FM broadcast band and proximity to channel 2, it would also have to have a very clean signal. Quite an order!

There simply weren't any commercial 6-meter amps that met my requirements. Ten-year old marginal commercial amps built before the "amplifier ban" days and 150-watt bricks were the sum total of selections available. Because most of the older units used a pi output circuit that just wouldn't provide enough harmonic rejection, they weren't clean and had second harmonics only 40 dB down. By today's standards, amps built for use above 50 MHz must exhibit second harmonic attenuation of at least 60 dB below the fundamental signal.

So I began to look for a design that would fit these parameters. After some research, it became clear that the pi-L output circuit, which provided a clean signal and sufficient harmonic suppression, would be the right way to go.

A 1000-watt dissipation tube was necessary. This meant that I'd have to use a pair of tubes such as those of the 8874 series or a single tube in the 8877 class. I ruled out glass envelope tubes because of high internal capacitances.

I chose the Eimac 3CX1000A7, a 1000-watt plate dissipation triode that's very similar to the 8877 except for a different socket. Parts are available from most parts supply houses. I used my junk box, RadioKit, Amp Supply, and a local parts supply house to obtain the necessary items.*

The amplifier is built on a 10 x 12 x 3-inch Bud chassis. (**fig. 1**). Aluminum panels (0.040 inch thick) are used for the chassis bottom cover, sides, and top plate. Aluminum "L" stock, available from most hardware stores, is used to hold the sides and top plate securely. All metalwork was done in my workshop using standard hand tools.

circuit description

The amplifier uses a grounded grid circuit with either the Eimac 3CX1000A7 or 8877, ceramic/metal triodes intended for linear service in the HF and VHF ranges. The amp provides the legal power output of 1500 watts PEP and CW service with no effort and requires a driver delivering between 50 and 80 watts at 50 MHz. This is well within the limits of multi-mode rigs.

With a plate voltage of 3000 volts at 0.8 amps the

By Lauren Libby, KX0O, 6166 Del Paz Drive, Colorado Springs, Colorado 80918

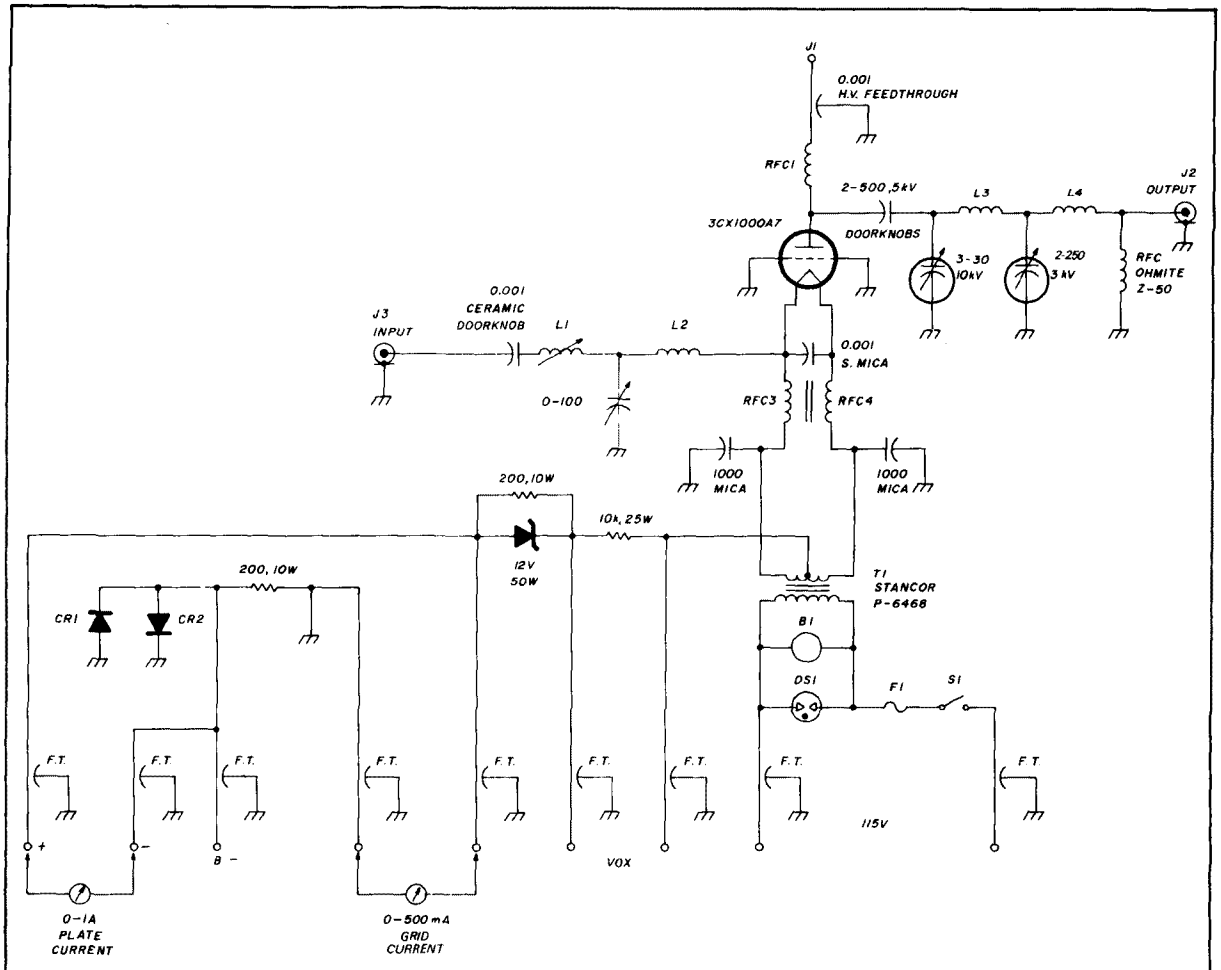


fig. 1. 6-meter amplifier schematic.

amplifier performs at 60 percent efficiency. The grid is grounded by means of the grid ring of the 3CX1000A7 socket providing a low-inductance path to ground. The amplifier is completely stable.

One notable feature of the amplifier is the use of a vacuum variable capacitor for tuning and loading controls in the plate circuit. The plate tuning capacitor is a 3-30 pF and the loading capacitor is a 0-250 pF. I used these units because they were available. The tuning capacitor should be a vacuum-variable because of the high voltages and currents present. An air-variable with large spacing could be used — just be sure not to exceed the voltage rating of the capacitor. The loading capacitor can be an air-variable that is of good RF transmitting design and rated for at least 2000 volts.

The pi-L circuit provides good harmonic suppression and broad frequency coverage. Most designs for the 6-meter band cover only the first couple of Megahertz of the band because of exotic designs using "floating turns" and expandable coil arrangements.

*RadioKit, Box 411H, Greenville, NH 03048; AmpSupply, 201 Snow Avenue, Raleigh, NC 27602

parts list

B1	*blower (Dayton 1C180)*
CR1,2	*1000-volt, 1-Amp diodes (Radio Shack)*
DS1	*110-volt lamp and holder*
FT	*0.001 500-volt feedthroughs*
J1	*Millen high-voltage connector*
J2	*"N"-type chassis connector*
J3	*BNC chassis connector*
L1	*6 turns, No. 18 on slug coil form available from Radio Kit, CTC 1538-4-3 or equivalent, spaced 7/8 inch*
L2	*6 turns No. 18, 1/2 inch diameter, 5/8 inch long, self-supporting*
L3	*3-1/4 turns, 1/4-inch copper tubing, 2-inch diameter 4 inches long*
L4	*8 turns, 3/16-inch copper tubing, 2-1/4 inches long, 3/4 inch diameter*
RFC1	*43 turns, No. 16 enameled, spacewound about 4-1/8 inches long on 3/4 inch diameter x 6-1/2 inch long teflon rod threaded to fit wire windings*
RFC2	*Ohmite Z-50*
RFC3,4	*30-Amp filament choke (available from Amp Supply) Bifilar wound on ferrite rod*
T1	*5-volt, 30-Amp filament transformer (Stancor) No. P-6468*

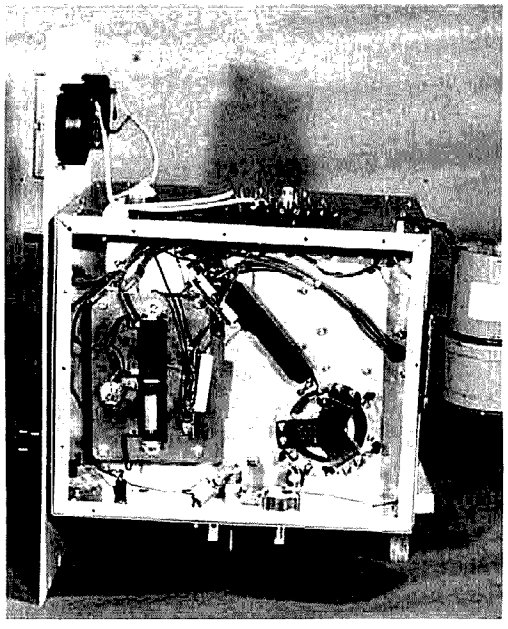


fig. 2. Grid compartment, bias supply, and input circuits. All connectors outside the compartment are made through feed-through capacitors. The input components are pictured in the lower center of the grid compartment.

These designs have been attempts to deal with high internal capacitances present in transmitting tubes. Only about 30 pF of capacitance are needed to resonate most circuits at 50 MHz, and many tubes have almost that much capacitance. So, in essence, the tube is a fixed tuning capacitor resulting in no tuning range for a variable capacitor.

This amplifier circuit provides for this condition by raising the *Q* of the circuit to allow for more capacitance tuning range. By increasing the *Q*, harmonic suppression is improved.

A 12-volt 50-watt zener diode is placed in series with the cathode return line to set the desired plate idling current and bias. The plate and grid circuits are metered in the cathode return lead.

the input circuit

The input circuit is a "T" design consisting of two coils and a shunt capacitor. One coil and capacitor are variable. With these two adjustments it's possible to cover a broad range of input impedances.

The controls are brought out the side of the unit for easy tuneup. Since the input circuit is a fairly broadband circuit, no tuning is needed once it's set for the first Megahertz of the band.

The input matching circuit can be seen in the lower center of the under-chassis photograph (fig. 2). The filament choke, a commercial unit obtained from Amp Supply, is capable of handling 30 amps of current at 5 volts.

A 1000-pF ceramic doorknob capacitor is used between the BNC input connector and the variable inductor in the input circuit. A transmitting mica would have worked just as well, as long as it was capable of handling 100 watts of drive power. Silver mica capacitors are used to isolate the filament line.

bias circuit

The bias circuit is built on a subassembly consisting of a brass plate and components. The subassembly is mounted on 1-inch spacers above the bottom of the chassis. The large resistors shown near the zener diode are suspended on a teflon rod hung from mounting supports.

plate circuit

Figure 3 shows the layout of the pi-L circuit. The plate choke is 43 turns of No. 16 gauge wire wound on a 3/4-inch diameter teflon rod measuring 6-1/2 inches long. I threaded the rod using an NC (National Course thread) die and then spacewound the coil in the threaded grooves. Half-inch copper strap is used to connect the plate circuit components. (This again is to accomodate the rigorous operating conditions.) Since the plate circuit has a high *Q*, large circulating currents are present. The vacuum variables are mounted with 0.060-inch aluminum "L" brackets that were fashioned with metal shears and bent to shape. The plate inductor L3 is wound from 1/4-inch copper tubing. The "L" portion of the circuit is mounted away from the loading capacitor at a 90-degree angle and is connected to a Z-50 (Ohmite) RF choke and antenna connector.

The layout is clean and easy to reproduce. Notice that an aluminum shield protects the filament transformer from stray RF.

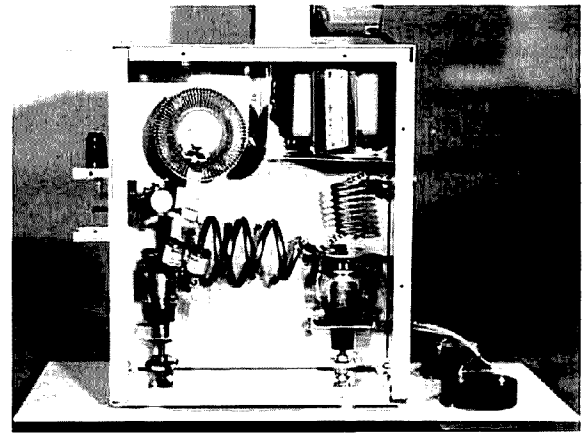


fig. 3. View of plate compartment. Note the location of the high voltage capacitor, plate choke, and plate circuit. This is an uncluttered layout and easy to reproduce.

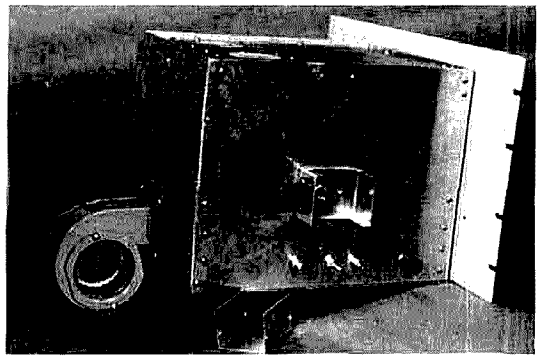


fig. 4. Rear view of completed unit. High-voltage connector and input tuning controls are pictured. The blower is positioned to pressurize the lower portion of the chassis.

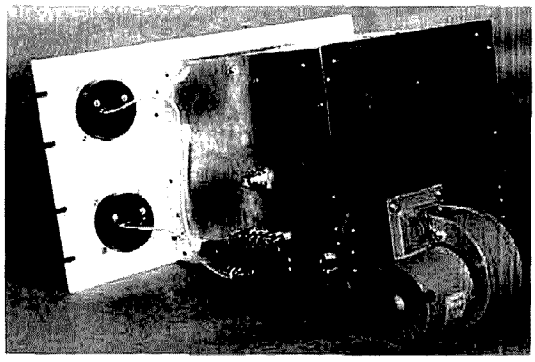


fig. 5. Completed rear view showing the feedthrough capacitors, output connector, meter mountings, and blower location.

The high-voltage connector is encased in an aluminum mini-box and connected to the high-voltage feedthrough capacitor. Safety is paramount when dealing with lethal voltages such as those used in this design (see fig. 4).

The tube chimney used with the 3CX1000A7 is a piece of teflon sheet (0.030 inch) cut to fit around the anode of the tube and fastened with teflon tape. Airflow is directed from below the chassis and through the socket, and is then exhausted through the tube and out the top of the amplifier (see fig. 5).

tube performance data

Either a 3CX1000A7 or an 8877 can be used in the circuit. The 3CX1000A7 uses either the Eimac SK-860 or SK-870 air system socket, a breechlock type. The 8877 uses a plug-in Johnson 122-247-202 socket.

Both tubes use 5 volts for the filament voltage. The primary difference is that the 3CX1000A7 requires 30 amps of current, since it has a directly heated filament, and the 8877 requries 10 amps.

Many operating parameters were tried with this piece of equipment. The most suitable ones for Amateur use are listed for the 3CX1000A7 below:

Plate Voltage	3000 Volts
Plate Current (Single Tone)	0.800 A
Plate Current (Idling)	0.125 A
Grid Voltage (Bias)	– 12 Volts
Grid Current (Full Drive)	0.200 A
Filament Voltage	5 Volts
Filament Current	30 Amperes
Power Input	2400 Watts
Power Output	1450 Watts
Drive Power	75 Watts

adjustment and tuneup

The completed amplifier is shown in **fig. 6**. Before applying any operating voltages, grid dip the input and output circuits to 50 MHz with the tube in the socket. This can be acccomplished by putting a 50-ohm load on both the input and output connectors. Tune the variable inductor and capacitor in the input circuit to resonate at 50 MHz. Note that this will have to be tuned again after the amplifier is operating; this will put it in the ballpark, however.

Grid dip the plate tuning circuit to 50 MHz without applying any voltages. You may have to expand or compress L3 to resonate the circuit. L4 should not require any adjustment and is not as critical as L3. However, it should resonate near 50 MHz.

Turn on the blower and apply filament voltage. Let the tube warm up, check the filament connections at the tube to ensure that a full 5 volts is present. Sometimes a voltage drop can occur across the filament choke. The tube should be within 0.25 volt either side of 5 volts.

Place a wattmeter in the input and output line of the amplifier. The input meter should be capable of handling at least 100 watts and the output meter should be capable of measuring 1500 watts output. I used a 2500-watt slug in my Bird meter. Tuneup was executed into a 1500-watt dummy load.

After the tube has warmed up, apply reduced plate voltage (about 2000 volts) and look for any evidence of arcing or other abnormal conditions. If everything appears to be operating normally, apply a small amount of drive and adjust the input circuit for minimum SWR. Do this quickly and with no more than 10 watts of drive.

Apply more drive gradually and adjust the plate tuning and loading controls for maximum output power. Look for any abnormal conditions. If none are observed, increase the drive and retune. If everything looks good, increase the plate voltage to 3000 volts and quickly tune the amplifier with about 50 watts of drive applied. Read the output power, compute the input power by multiplying plate voltage by the plate

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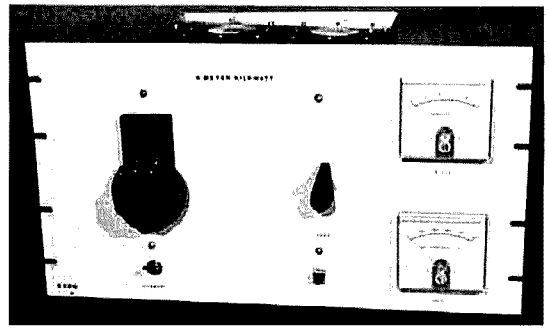


fig. 6. Front view of the "Colorado Kilowatt."

current and compare it with the output power measured. This will give you an idea of the plate circuit efficiency. If everything appears normal, increase the drive to 80 watts, set the parameters suggested in this article, and you'll have a commanding signal on the 6-meter band.

final comments

Several lessons were learned in designing and building this amplifier. One was not to hook the connection of the high-voltage strap going to the tube to a bolt running through the teflon rod of the RF plate choke. During tuneup I had flaming teflon rod as a result of all the circulating currents running through a 6-32 screw. Make sure the connection of the tube strap is on the same side as the one to which the choke winding is attached.

Before you begin construction of this amplifier, I'd suggest you do some reading: Bob Sutherland, W6PO's "Two-Kilowatt Amplifier for Six Meters," which appeared in *ham radio* in February, 1971, and Bill Orr's *Radio Handbook*, regarding the design parameters of the pi-L circuit, are "must" reading.

Although you don't have to be an electrical engineer to build this amplifier (I'm an economist), it's not a project for the first-time builder. But if you have some experience, you may want to build it and enjoy the operating possibilities it affords. As you build, it may help to keep in mind something I've found to be true over several years of building amplifiers: for cleaner signals and years of trouble-free operation, it's best to over-build.

Because many good high-voltage circuits have already been published, I haven't included a power supply circuit with this article. Just be sure to make your power supply husky; if you have trouble lifting it, it's big enough.

acknowledgments

Many thanks to Ray Uberecken, AA0L, and Hal Bergeson, W0MXY, for their encouragement and help. Special thanks to Reed Brandon of Eimac for hours of help in the initial design of the circuit.

ham radio

Inexpensive IC makes front end "uncrunchable"

a broadband amplifier-attenuator

The step attenuator is an important RF building block. On the workbench it can be used for testing the performance of amplifiers, mixers, or entire receiver systems; they're particularly useful for intermodulation tests. In test equipment they're useful for level references in spectrum analyzers and other instruments requiring wide dynamic range RF level measurements.

It's unfortunate, however, that step attenuators aren't often used in Amateur Radio receiver front ends. Anyone who's ever operated contests with a low dynamic range receiver knows the disastrous consequences of receiver overload and intermodulation interference; it's usually the third-order intermod products that cause problems in receivers.[1,2]

The equations that define the third-order product levels indicate that if you reduce the input signal levels by 1 dB, the third-order products will decrease by 3 dB. Thus if you "switch in" 5 dB of attenuation at the front end, the intermod products will go down 15 dB. With multiple strong signals coming through the receiver bandpass, the step attenuator can indeed be a useful feature; under contest conditions, for example, there's no need for an expensive receiver to become less useful than it might be when the addition of a few resistors and switches could have been used to eliminate the problem (**fig. 1**).

Another problem arises during low signal level conditions, particularly at higher frequencies. Since the signal-to-noise ratio and "noise figure" of the receiver are largely determined by the front end, using a low-noise amplifier will improve ultimate sensitivity. The typical HF receiver front end consists of a tuned amplifier complemented with some type of gain-control circuitry. This gain control can be automatic (AGC) or manual (RF). Front-end design is complicated by the tuned circuits necessary for image rejection if a low IF is chosen (<20 MHz). However, the trend in modern Amateur Radio receiver design is to first "up-convert" to a relatively high IF (typically 45 MHz). This technique places the image well up into the VHF spectrum and eases the front-end filtering requirements.

broadband constant gain and impedance

Some type of bandpass and impedance matching circuit is still required as an interface between the antenna and the usual discrete amplifier device. Unfortunately the input and output impedances of the familiar bipolar and FET devices do not match the universal 50-ohm standards.

The front-end amplifier problem can be greatly reduced by designing with a new broadband amplifier, the Signetics NE5025 (**fig. 2**). Priced at about $1.50, the 5205 is a 20-dB gain block using a multiple feedback scheme that normalizes the input impedance over a very wide range. Both gain and impedance show only slight variation from DC to 650 MHz! The input impedance of the amplifier is simply "imaged" by the output termination; that is, if you terminate the output in 50 ohms, the input impedance will also be 50 ohms. This particular characteristic, in monolithic form, should be of interest to RF experimenters.

The 5025 also works in systems of 75 ohms and higher impedance. JFET double-balanced mixers (for example, the Siliconix U350) and the new ultra-high performance DMOS quad rings (such as the Siliconix Si8901) have a relatively low but constant input impedance, typically 12 ohms. Thus the 5205 can feed an 8901 through a 4:1 broadband balun for broadband front-end operation up to a few hundred MHz. This example completes a high-performance radio front-end concept free of resonant tuned circuits.

By Robert J. Zavrel, Jr., W7SX, 707 Borello, Mountain View, California 94041

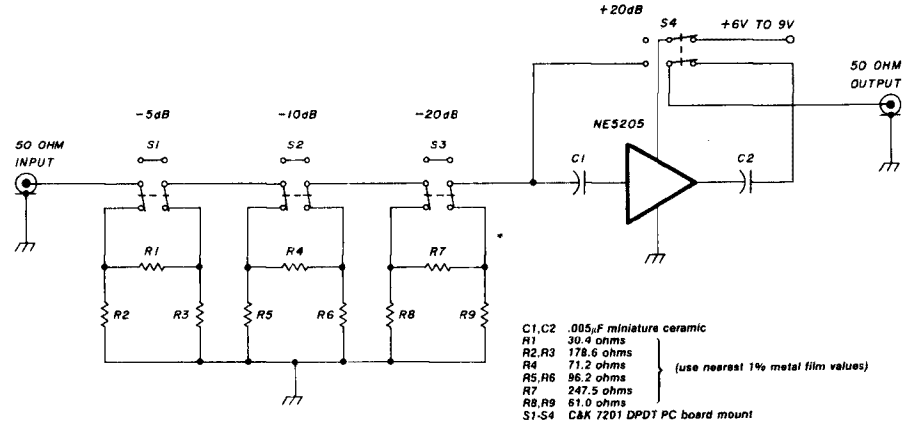
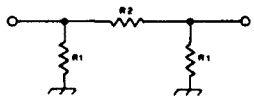


fig. 1. 35-db step attenuator and gain stage improve intermod performance.

Table 1. Pi-network resistive attenuator (50 ohms).

dB Atten	R1 (ohms)	R2 (ohms)
1	870.0	5.8
2	436.0	11.6
3	292.0	17.6
4	221.0	23.8
5	178.6	30.4
6	150.5	37.3
7	130.7	44.8
8	116.0	52.8
9	105.0	61.6
10	96.2	71.2
11	89.2	81.6
12	83.5	93.2
13	78.8	106.0
14	74.9	120.3
15	71.6	136.1
16	68.8	153.8
17	66.4	173.4
18	64.4	195.4
19	62.6	220.0
20	61.0	247.5
21	59.7	278.2
22	58.6	312.7
23	57.6	351.9
24	56.7	394.6
25	56.0	443.1
30	53.2	789.7
35	51.8	1405.4
40	51.0	2500.0
45	50.5	4446.0
50	50.3	7905.6
55	50.2	14,058.0
60	50.1	25,000.0



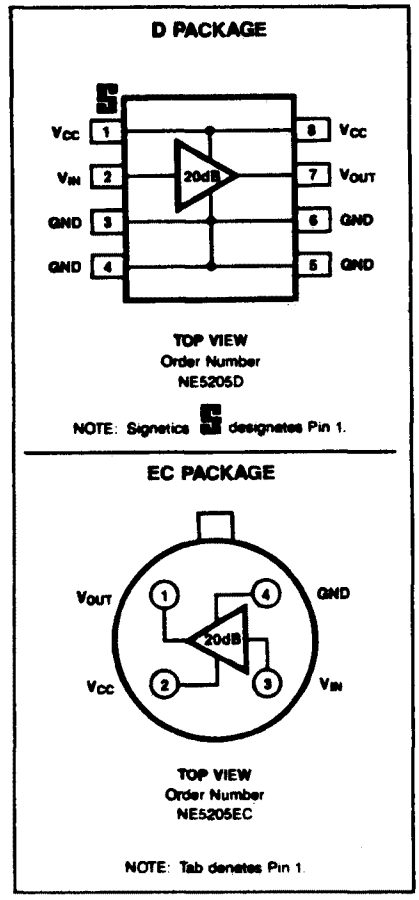
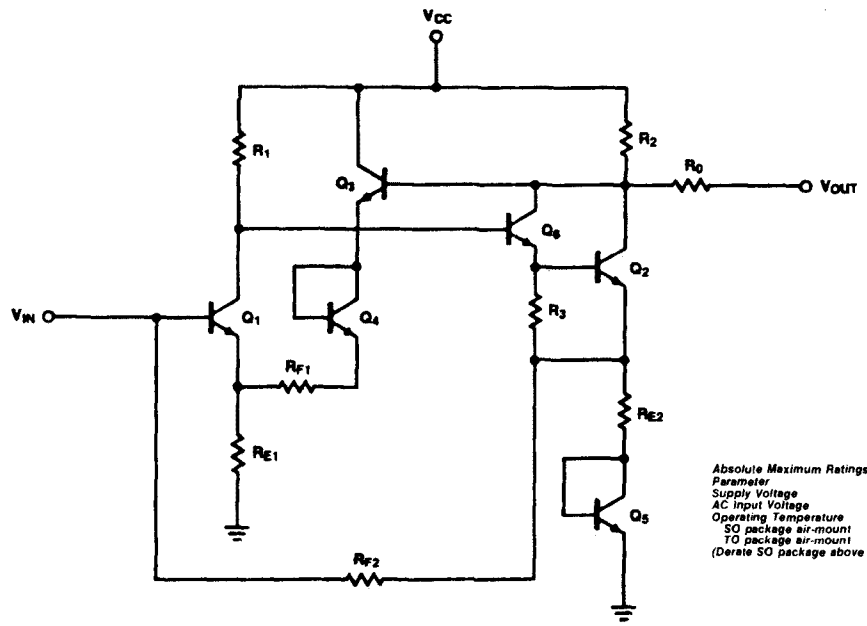


fig. 2. Pin configuration of Signetics NE5205.



Absolute Maximum Ratings		
Parameter	Rating	Units
Supply Voltage	9	V
AC input Voltage	5	V-PP
Operating Temperature		
SO package air-mount	-55 to +85	°C
TO package air-mount	-55 to +125	°C
(Derate SO package above 6V)		

fig. 3. Equivalent internal schematic for NE5205.

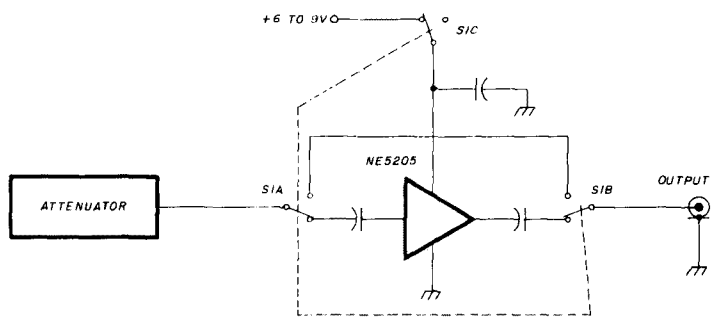


fig. 4. Use of multi-wafer ceramic miniature switch for NE5205 in/out switch.

Some filtering is still required, however. The use of upconversion techniques allows installation of a simple multi-pole 35- or 40-MHz low-pass filter ahead of the attenuator for image rejection. Inclusion of a broadcast band high-pass filter is also desirable for strong AM signal rejection.

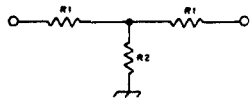
overcoming circuit constraints

The circuit shown in **fig. 1** allows both attenuation and amplification. Although using mini-toggle switches limits the frequency response to about 350 MHz, they are easy to mount, inexpensive, and rugged. This circuit is practical for the 160- through 2-meter bands.

A few precautions are necessary when using the 5205. It's usually easy to use if good RF techniques (i.e., providing adequate grounding and keeping the leads as short as possible) are employed. Care must be taken, however, when switching both the input and output circuits with a common switch assembly. The switch forms a capacitor between the input and output, while the switch leads and circuit board etching form a strip inductance. The resulting series LC circuit in the feedback loop will cause the amplifier to oscillate at a VHF or UHF frequency. The circuit shown oscillates at about 600 MHz, but the level is well below saturation, and the spectrum is very clean below 400 MHz. All oscillations can be eliminated by

Table 2. T-network resistive attenuator (50 ohms).

dB Atten	R1 (ohms)	R2 (ohms)
1	2.9	433.3
2	5.7	215.2
3	8.5	141.9
4	11.3	104.8
5	14.0	82.2
6	16.6	66.9
7	19.0	55.8
8	21.5	47.3
9	23.8	40.6
10	26.0	35.0
11	28.0	30.6
12	30.0	26.8
13	31.7	23.5
14	33.3	20.8
15	35.0	18.4
16	36.3	16.2
17	37.6	14.4
18	38.8	12.8
19	40.0	11.4
20	41.0	10.0
21	41.8	9.0
22	42.6	8.0
23	43.4	7.1
24	44.0	6.3
25	44.7	5.6
30	47.0	3.2
35	48.2	1.8
40	49.0	1.0
45	49.4	0.56
50	49.7	0.32
55	49.8	0.18
60	49.9	0.10



using a miniature ceramic rotary switch configured as **fig. 4**. The goal is to raise the feedback resonant frequency above the gain cut-off point of the 5205. This is accomplished by reducing switch capacitance and lead inductances. The 5205 can also be left "on" at all times, thus eliminating the oscillation problem.

construction techniques

The test unit was built on a 1 x 3-inch PC board with double-sided ground planes. The PC board can be mounted in a die-cast minibox that also holds the input and output sockets (BNC, N, or UHF) and the mini-toggle switches. The 5205 is available in standard N, surface mount, and TO-46 packages. If the TO-46 is used, the package should be mounted in a snug hole and the case soldered to the ground plane. Miniature ceramic input and output capacitors should be soldered directly to the 5205 leads. Good UHF design techniques must be employed even if the unit is used only at HF.

good intermod performance

Attenuation in 5-dB steps was chosen. This is somewhat arbitrary; other steps can be selected. **Tables 1** and **2** are included for convenience.[3] This circuit, shown in **fig. 1**, allows gains of −35 to +20 dB in 5-dB steps, which should be adequate for any HF receiver requirements. The attenuators should precede the amplifier to maximize amplifier linearity.

The noise figure of the 5205 is less than 6 dB at 100 MHz for 50-ohm systems, dropping to 4.8 dB at 75 ohms. The third-order output intercept point is +17 dBm. This indicates that it will take two signals 33 dB above S9 to produce a third-order product at an S1 level.[4] With only 5 dB of attenuation inserted, two signals must be 48 dB above S9 to produce an S1 level third-order intermod. This is indeed a strong front end!

summary

This simple project can greatly enhance the utility of a step attenuator on the test bench. In a receiver front end the NE5205 can eliminate bandpass filters while exceeding the intermodulation characteristics of many common mixers. Finally, when combined with a very strong mixer (such as the Si8901) an entire broadband, high-performance receiver system becomes possible.

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3. 1986 *ARRL Handbook*, American Radio Relay League, 1986, Chapter 25.
4. Robert J. Zavrel, Jr., W7SX, "A Calibrated S-Meter", *ham radio*, January, 1986, page 23.

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ham radio TECHNIQUES

Bill Orr W6SAI

a summer miscellany

Summer's here! Even though the sunspot cycle doesn't promise much DX, these *are* the months for antenna work. Here are some interesting designs you may want to try.

7-MHz sloping antenna

Many Amateurs have had success with the delta loop on the lower frequency bands. I've had queries about the effect of mounting the loop at an angle in order to conserve pole height. Does tilting the plane of the loop impede its fine DX record? I didn't know the answer to this question, but Kjell, SM6CTO, decided to try out the sloping loop shown in **fig. 1**. He found out that it worked very well; while he couldn't make a direct comparison with a vertical loop, he found it to be "a useful DX performer." (The antenna was described in the September, 1985, issue of *Radio Communication*, a publication of the Radio Society of Great Britain.)*

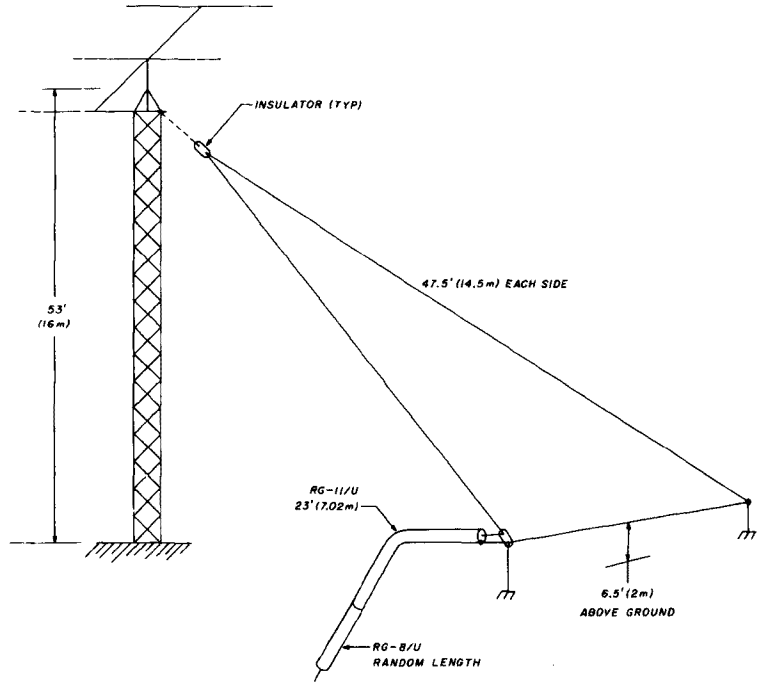


fig. 1. 7-MHz sloping delta loop at SM6CTO.

broadband 80-meter dipole

There's been a lot of sound and fury about the so-called broadband dipole for the 80-meter band (3.5 to 4.0 MHz). Many designs have been published and some of them work. One design that does work was recently described by Malcolm Johnson, VK6LC, in the November, 1985, issue of *Amateur Radio*, published by the Wireless Institute of Australia.* The layout is shown in **fig. 2**.

In brief, it's a cage dipole about 102 feet long, fed by a matching transformer and a 1-to-1 balun. The SWR on a 50-ohm transmission line across the 3.5- to 3.8- MHz range is less than 1.8-to-1, with a minimum SWR of 1.3-to-1 at the design frequency of 3.65 MHz. It's ideally suited for use with a solid-state transmitter. Less than half a wavelength long, the five-wire cage exhibits a capacitive reactance at the feedpoint that's transformed by the parallel line transformer to become slightly inductive. A shunt capacitor at the feedpoint parallel-resonates with the inductive reactance to aid in improving antenna bandwidth response.

The dipole is erected in inverted-V fashion, with the center apex at 36 feet above ground.

Malcolm used a four-way splice connector called a "Queblock" to make his insulated spreaders. In any event, it seems that a little ingenuity will produce a spreader that will do the job. Six spreaders are required.

The two-wire transformer is about 0.08 wavelength long (20.7 feet) and has an impedance of 300 ohms. VK6LC used a home-made line, but I think that 300-ohm ribbon line could be used for low-power operation, provided the velocity facor of the line is taken into account. The shunt capacitor can be a high-voltage disc unit or mica capacitor of 1 kV or better, for powers up to several hundred watts.

In order to obtain an accurate SWR profile of the antenna, VK6LC used a half-wave long transmission line.

*Walter Schulz, K3OQF, "Designing A Vertical Antenna," *QST*, September, 1978, page 19.

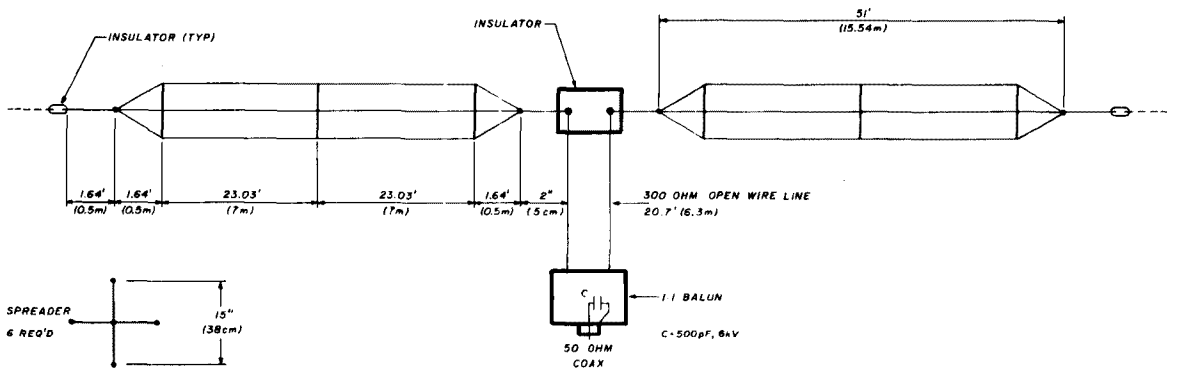


fig. 2. The 80-meter cape diode at VK6LC. A 500pF capacitor is placed across at the balun.

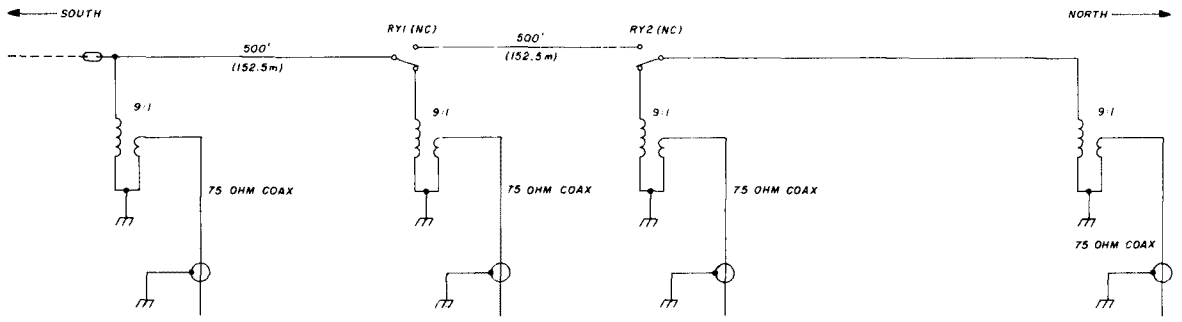


fig. 3. Variable length Beverage antenna at WB3GCG. Relay selection at operating position permits choice of 500, 1000 or 2000 foot wire. Power to relays runs through coax line.

The center resonant frequency of the antenna can be easily changed by adding or subtracting 5 inches of the 300-ohm line for each 20 kHz of desired frequency change.

Results? Plenty of DX worked, including such juicy examples as A71AD, ON5YA, YU4EBL, H44IA, WA6SLO, HA5XW and others — all with 100 watts! That's not bad for any antenna on the QRM-filled 80-meter band!

all you need is . . .

. . . a bunch of 200-foot towers! I just received a letter from Willy, WB3GCG, who has changed occupations and moved to a new QTH. He's had some big 160-meter antennas in the past and is looking forward to great DX days ahead. On 20 acres of land, he's erecting three 200-foot high steel towers arranged in a triangle, approximately 500 feet apart. The towers will support the ends of several Beverage antennas, each 2000 feet long. Relays will be placed along the Beverage wires so the wire can be either 500, 1000, or 2000 feet long. When he's finished, Willy will have eight Beverage antennas. The feed system for the antennas is as shown in **fig. 3**. Each wire section is terminated in a 9-to-1 transformer. Coax lines run to the operating position; the wire lengths and terminating resistors are selected through a control box.

But that isn't all. Willy uses separate antennas for transmitting and receiving. For transmitting, he favors a large vertically polarized loop (**fig. 4**). Similar loops are used as directors and reflectors. He can switch to horizontal polarization by means of a vacuum relay located at the apex of the loop.

A large number of loops are strung on catenary cables between the towers. Willy aims to have three separate 160-meter delta loop beams, each consisting of back-to-back four-element parasitic arrays with a common driven element (see **fig. 5**), strung between the towers! With the use of multiple feedlines and vacuum relays, each delta beam can be quickly reversed in direction. It takes a total of seven loops and 3500 feet of No.14 four-strand, vinyl-covered wire for one beam!

In closing, Willy remarks that the

delta loop beam really works. All of this reminds me of the song, "Will everyone here/kindly step to the rear/and let a winner lead the way!" Good luck, WB3GCG, and may the wintry winds be kind to your antennas.

the "rinky-dink" antenna at K6FD

Yes, that's what Ray, K6FD, calls his tuned 40-meter V-doublet (fig. **6**). He uses it on the 30, 40, 80, and 160-meter bands. The antenna is "tuned" from the operating position with a T-network. The included angle between the legs is about 75 degrees, and the wires slope down slightly towards the ground. The antenna is decoupled from the feedline with a simple 14-turn coil, 3 inches in diameter, of small RG-141/U coax. The remainder of the feedline is RG-58/U.

The simple network allows unity SWR to be achieved on all bands. The coax feedline is about 100 feet long. Ray says that with appropriate coils in the network, the system will tune through the broadcast band.

Ray recommends this antenna to those operators who would like to try 80- and 160-meter operation but haven't done so because they haven't room for large antennas.

a cheap, top-loaded antenna for 160 meters

Arne, K0AS, is lucky in that he has a number of tall trees on his property. He wanted to work the low end of 160 meters and found design information for a top-loaded vertical antenna in an article in an old copy of *QST*. Following that design, he used a 50-foot high wire with plenty of top loading. The top loading disk, a rectangle in this case covering 480 square feet, is made of No. 10 copperweld wire. An L-network is placed at the base of the antenna to match it to a 50-ohm coax line. Beneath the antenna are 50 radials, each 125 feet long. A sketch of the antenna and the matching network is given in **fig. 7**. The feedpoint resistance of the antenna is about 18 ohms.

The resonant frequency of the antenna can be raised to about 1850 kHz by manipulating the switches in the base network. At this frequency, the operating bandwidth between the 2-to-1 SWR points on the feedline is about 60 kHz.

Arne says the antenna is "quite inexpensive and the results are most gratifying." You couldn't ask for much more than that.

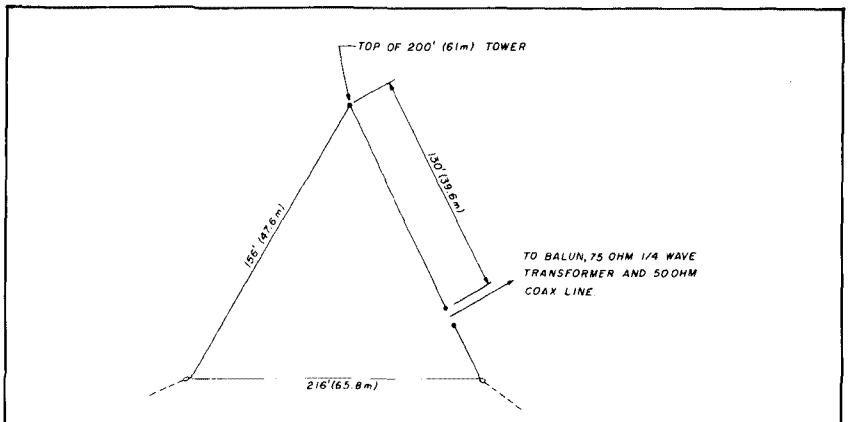


fig. 4. Vertically polarized delta loop for 160 meters in use at WB3GCG.

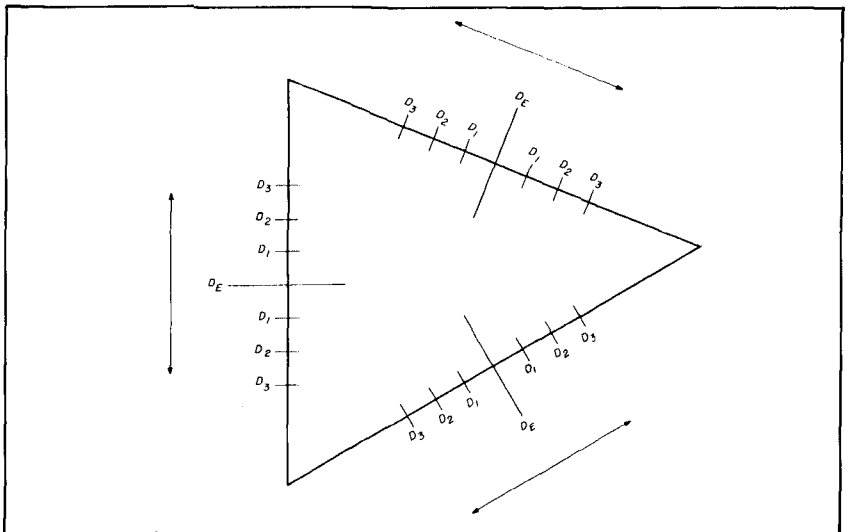


fig. 5. WB3GCG's delta loop consists of seven 500-foot loops of wire, for a total of 3500 feet.

another nifty low-band antenna

Bob, K9EVI, wanted an antenna that would cover 160, 80, 40, and 30 meters and fit in his back lot. What to do? He had a tower that supported a tribander for 20, 15, and 10 meters — now he needed an antenna for the low bands.

His solution was to use a center-fed dipole, 45 feet on a leg. The antenna is slung, in inverted-V fashion, from the top of his 45-foot tower. The ends are about 15 feet above ground level. Bob uses a 600-ohm, open wire feedline about 45 feet long that runs down to his station and into a simple matching network (**fig. 8**) for 80- and 160-meter operation. For the other bands, he uses an old Johnson "Matchbox" tuner. On 80 and 160 meters, the antenna feeders are tied together and the system works as a

top-loaded vertical antenna, with the feeders doing the major portion of radiation.

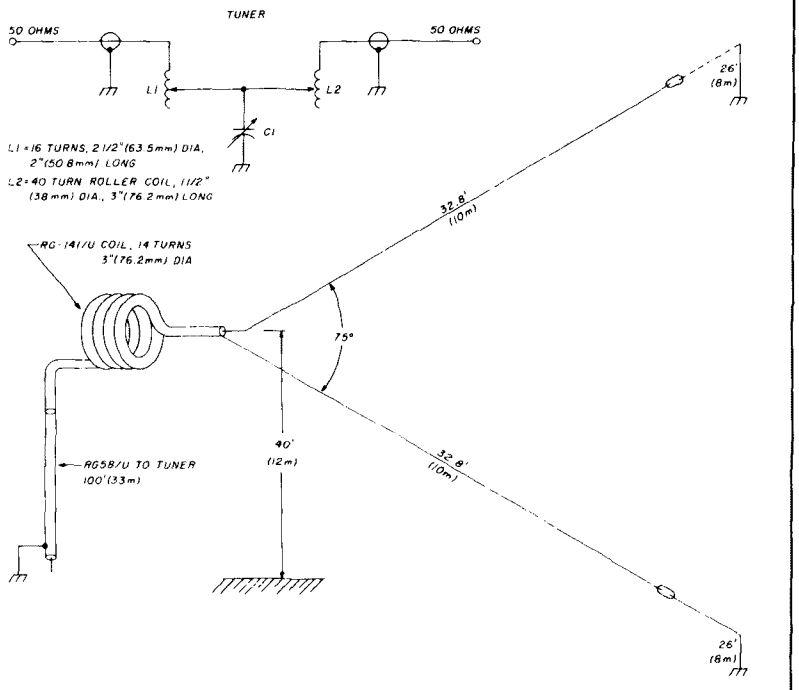


fig. 6. Multiband antenna and T-tuner at K6FD.

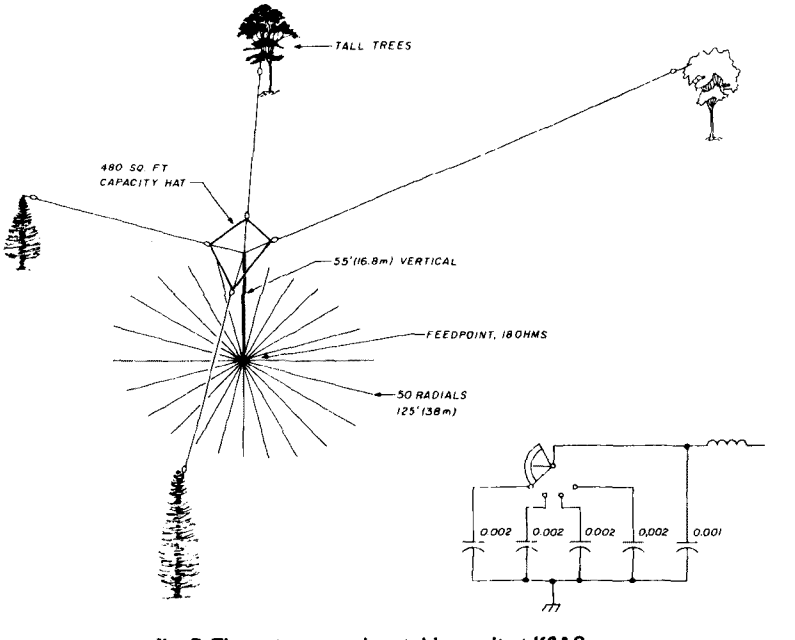


fig. 7. The antenna and matching unit at KØAS.

the first radio handbook

The Wireless Experimenter

proclaimed it "the publishing event of the year." Experimenters in the United States and Canada bought out the first edition even before it was on the presses. And the General Electric Company bought 500 copies sight-unseen.

Starting with Marconi's famous experiments at the turn of the century, the advance of wireless (i.e., radio) rushed ahead in Europe. Following in the footsteps of Marconi, experimenters and engineers in Germany quickly advanced the radio art. But very little of this advanced knowledge crossed the Atlantic to the United States. Hearsay, letters, and months-old magazines were the only sources of information available to experimenters and Radio Amateurs in the United States.

There were, however, rumors of a wonderful handbook, written in German, entitled *Lehrbuch fur Dratlose Telegraphie* (*Textbook for Wireless Telegraphy*). A few copies could be found in college libraries, but the information was obviously accessible only to those who had a good command of technical German.

The 1912 version of the book — written in 1905, published in Braunschweig in 1908, and revised in 1912 — was translated by A.E. Seelig of Wellsville, New York, and published by McGraw-Hill in the fall of 1915. And what a wonderful book it was! Written by Professor Jonathan Zenneck of the Technical High School of Danzig, the 440-page hardbound book clearly defined and explained the wonderful world of wireless. The book bristled with photographs of arc and spark transmitters and various types of crystal, carborundum, magnetic, and diode receivers. Much attention was given to antenna design and installation. Long-range propagation via the ionosphere was discussed along with gray-line propagation at sunrise and sunset.

Interestingly, comprehensive data was provided on voice transmission using an arc transmitter, a technique seemingly known to only a handful of experimenters.

This beautifully illustrated book, printed on cream-colored heavy paper, made Professor Zenneck an instant celebrity in the United States.

the doctor is interned

Shortly before World War 1, Dr. Zenneck came to the United States. Welcomed with a dinner at the Institute of Radio Engineers in New York, he visited many radio laboratories, as well as the German-owned radio station at Sayville, New York.

On July 6, 1917 the wireless world was stunned by a headline in the New York Times: ARREST GERMAN RADIO EXPERT AND TAKE HIM TO ELLIS ISLAND. The Department of Justice had determined that Dr. Zenneck was "one of the most dangerous German subjects in this country," and claimed that he had participated in "the memorable German drive through Belgium" and had later "falsified his passports" to gain admission to the United States **(fig. 9)**.

Were these accusations true? No matter. Dr. Zenneck was interned at Ellis Island for the duration of the war. Released at the end of the war, he reportedly enjoyed a reunion dinner with his colleagues at the Institute of Radio Engineers, but thereafter disappeared from history.

By that time the glamour of "Zenneck" had faded. Great strides in radio communication had been made during the war period and the vacuum tube was supplanting the spark, the arc, and the crystal detector. DeForest's audion, which originally rated a half-page in "Zenneck," was now supreme and large portions of the first Radio Handbook were swept into the dustbin of history.

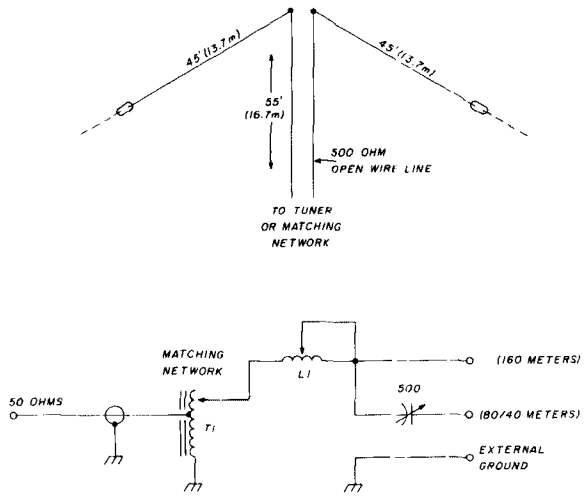


fig. 8. K9EVI ties feeders together on his HF center-fed antenna and operates it on 160, 80, and 40 meters.

ARREST GERMAN RADIO EXPERT AND TAKE HIM TO ELLIS ISLAND

Professor Zenneck Is to Be Interned During War By Order From Washington.

By the United Press.

Trenton, N. J., July 6.—Acting under special orders from the Department of Justice at Washington, United State Deputy Marshal Linford Denny, of this city, late last night placed Prof. Jonathan Zenneck, German radio expert, under arrest and took him to Ellis Island, where he will be interned for the duration of the war.

The Department of Justice is said to look upon Professor Zenneck as one of the most dangerous German subjects in this country. Germany looks upon him as one of her most skilled wireless telegraph experts, and he came to this country especially to direct German wireless activities. For a long time he was in charge of the German radio station at Sayville. Of late he has been living at Boonton, at which point he was arrested by Deputy Denny.

Before coming to America Professor Zenneck served in the German Army in an official capacity. He participated in the memorable German drive through Belgium and later by falsifying his passports gained admission to the United States.

4

fig. 9. *The New York Times* **reported arrest and detention of Professor Zenneck.**

"Zenneck" today

Looking at yesterday through the eyes of today is an instructive experience. Zenneck had a firm grasp of the basic principles of wireless. Much of his book is fundamentally correct and the work stands as a monument to the author and the many prominent scientists who assisted him. Copies of the first edition of *Wireless Telegraphy* (1915) are very rare and hard to locate. (I found mine in a second-hand bookstore.) It is prudent for connoisseurs of the radio art to be alert for a copy of this work because it represents the very foundation of radio handbooks today.

A belated — and much-deserved — congratulations to Professor Jonathan Zenneck for a job well done!

2-meter EME directory

The 16-page directory of active 2-meter "moonbounce" stations worldwide, listing each station's name, QTH, and equipment used, has been reprinted. For a copy, write to me at EIMAC, 301 Industrial Way, San Carlos, California 94070. Enclose five first-class stamps (or five IRCs) for postage.

ham radio

extending receive coverage
for the IC-02 and IC-04

Simple, reversible modifications let you monitor public service frequencies

The ICOM IC-02 (2 meters) and IC-04 (70 cm) handhelds are a natural step up from the earlier IC-2 and IC-4. Microprocessor controlled, these newer models feature a liquid crystal display, an S-meter, a built-in CTCSS (continuous tone-coded squelch system), and scanning capability, all in a water-resistant package that accepts accessories designed for the IC-2 series.

The United States version of the IC-02 transceives from 140 to 149.995 MHz; the IC-04 will transceive on the upper 10 MHz of the 70-cm band from 440 to 449.995 MHz. With the simple modifications described in this article, the IC-02 can be adapted to receive from 140 to approximately 163 MHz; the IC-04 can be modified to receive up to about 465 MHz. Transmit capability is unaffected, and any frequency entered will be properly stored in memory. Because this isn't a permanent modification, each rig can easily be returned to its original state.

Extending the receive capability of either unit will let you listen to the public service frequencies (police, fire, and radiotelephone, for example) located just above the 2-meter and 70-cm bands. Modifying the IC-02 also permits reception of National Weather Service broadcasts at 162.40, 162.475, or 162.55 MHz in most metropolitan areas. Receive sensitivity for frequencies this far removed from the 2-meter band is reduced, but should be adequate for most applications.

the microprocessor

The IC-02 and the IC-04 are each controlled by a Hitachi HD44795 80-pin microprocessor which operates the display, keyboard, and frequency synthesizer. The processor operating environment is determined by an initialization matrix (**figs. 1** and **2**) consisting of four output and four input lines connected to each other in a specific pattern with isolating diodes. This pattern determines processor reset frequency, upper and lower band limits, step size, method of frequency entry from the keyboard, and other functions. The modification procedure described here changes the initialization matrix to remove the receive frequency limits as determined by the microprocessor. Receiver coverage, however, is still limited by the tuning range of the synthesizer VCO and by receiver sensitivity.

disassembling the units

Before you start, make sure you have a soldering iron with a *very* fine tip; you'll be working with extremely small components. [If you've never worked with such small components, in such tightly confined space, you may want to practice first. — Ed.]

Begin by disassembling the battery pack and antenna from the transceiver. Then remove the battery mounting plate and locking lever assembly. Unfasten the five screws holding the metal back cover to the rig and take off the cover, using care not to damage the "O" rings that seal the back cover to the rest of the case. (On the IC-04 only: remove the large metal shield covering the PLL circuitry.) Now separate the chassis and top section and place them next to the front case; note the ribbon cable connecting the chassis to the front section. Undo the two screws on the DTMF/CTCSS board and set the board off to the other side of the front case (02AT/04AT only). The unit is now ready for modification.

By Robert K. Morrow, Jr., WB6GTM, RR 1, Box 6F, Flora, Indiana 46929

initialization matrix

The microprocessor board is identical in the IC-02 and IC-04 except for the diodes attached to the initialization matrix area located at the top left corner of the board (**fig. 3**). We'll be working with the portion of the matrix connecting microprocessor lines R10-R13 to R20 and R21.

ICOM uses very small three-pin chip diodes — type A3, D3, or E3 (indicated on the diode itself) — to pro-

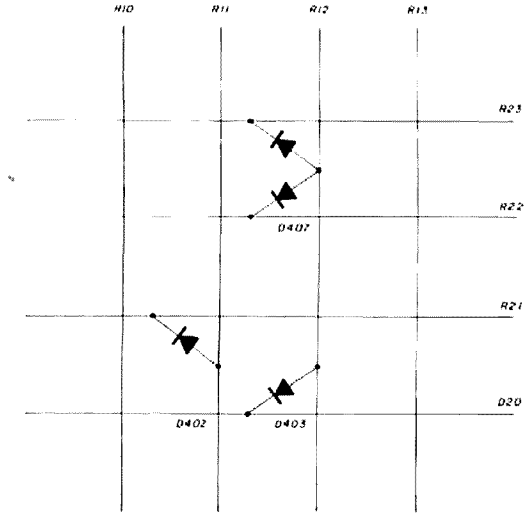


fig. 1. Initialization matrix for the factory IC-02. Removing D402, which is a type D3 located at position D2 in fig. 3, will remove the upper receive frequency limit of the rig.

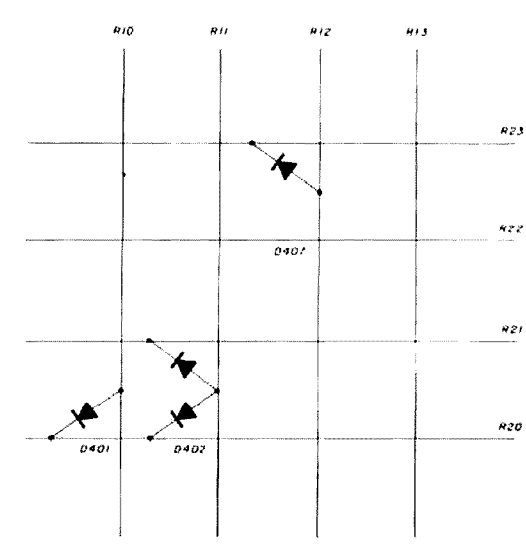


fig. 2. Initialization matrix for the factory IC-04. Removing D402 and connecting a 1N914-type silicon diode between R11 (anodel and R20 (cathode) will remove the upper receive frequency limit of the rig.

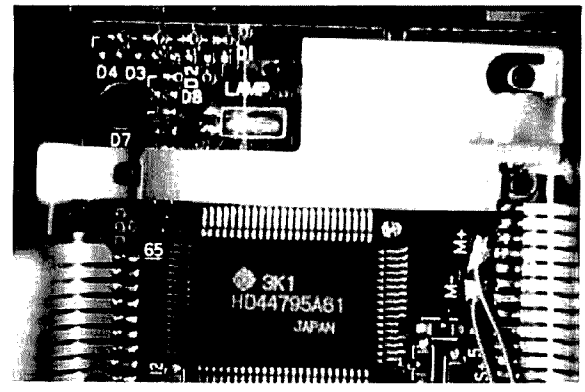


Fig. 3. Initialization diodes can be located at positions D1-D4, which are found at the top left corner of the microprocessor circuit board. The factory diodes are surface mounted on the triangular pads at D2 and D3 for the IC-02, and at D1 and D2 for the IC-04. The IC-02 is modified by simply removing the diode at position D2. To modify the IC-04, remove the diode at position D2 and replace it with a 1N914-type connected from the top pad (anode) to the lower right pad (cathode).

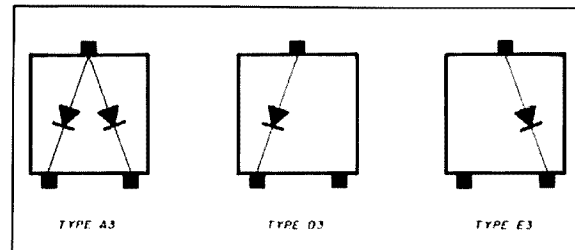


fig. 4. These chip diodes are used to initialize the microprocessor on the IC-02/04. Type A3, D3, E3, or no diode at all will be found at various points in the initialization matrix.

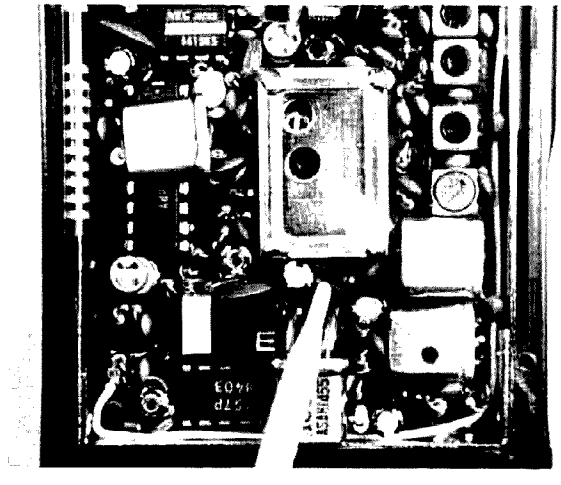


fig. 5. The IC-02 is shown completely reassembled except for the back panel. The test clip is connected to the VCO test point, which is the only vertically mounted resistor in the area with a bare upper lead. The arrow on the VCO shield points to the coil which should be adjusted for 0.75 VDC at the test point.

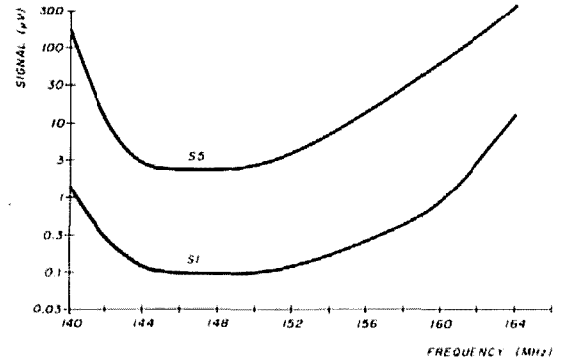


fig. 6. Graph of sensitivity vs frequency for the IC-02. S1 is also called the squelch sensitivity, and can be considered the minimum useable signal level. S5 is the signal level required to activate five pairs of LCD bars on the transceiver S-meter.

gram the initialization matrix; in this way component count and size are minimized (**fig. 4**). Do not confuse diode *types* with their *locations*, (D1, D2, D3, D4) printed directly on the microprocessor circuit board. These locations are labeled D401, D402, D403, and D404 on the ICOM schematic, but only for the diodes which are actually present.

IC-02 modification

The IC-02 is factory-wired with a type D3 diode at position D2 and a type E3 diode at position D3. The diode at position D2 restricts the receiver to the 140-151.995 MHz range. Remove this diode with care.

Now reassemble the IC-02, but leave the back cover off. Connect a high-impedance voltmeter to the VCO test point shown in **fig. 5**, attach the battery pack, and turn on the rig. Select 144.0 MHz from the keyboard.*

Adjust the coil in the VCO shield (lower hole) for a reading of 0.75 VDC. Be sure to use a properly fitting plastic alignment tool (or try whittling down a standard one with a razor blade). This alignment procedure maximizes the receive frequency range by setting the VCO near its lower voltage limit at the lowest transceive frequency desired. The receiver will now operate from slightly below 140 MHz to slightly above 163 MHz, and the transmitter will generate a clean signal throughout the 2-meter band. **Figure 6** shows receiver sensitivity for the entire VCO range. Now reinstall the back cover.

IC-04 modification

The IC-04 is factory-wired with a type E3 diode at position D1 and a type A3 diode at position D2. Carefully remove the diode at position D2 and replace it with a 1N914-type silicon diode connected from the top pad at D2 (anode) to the bottom right pad at D2 (cathode). This removes the receive limit portion of the matrix but still initializes the processor at 440 MHz.

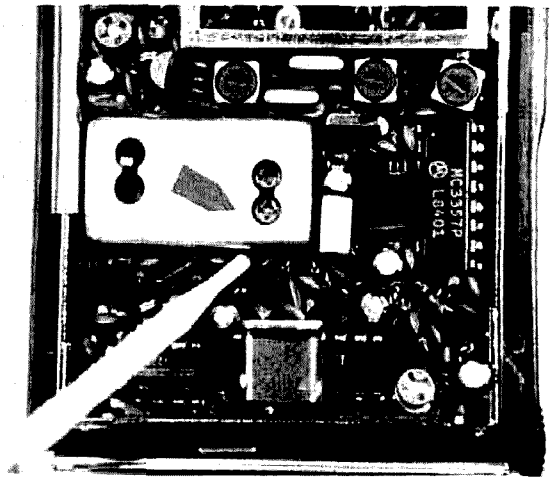


fig. 7. The IC-04 is shown completely reassembled except for the back panel and shield, and the metal tape has been removed from the VCO can. The test clip is connected to the VCO test point which, like the IC-02, is the only vertically mounted resistor in the area with a bare upper lead. The arrow on the VCO shield points to the variable capacitor which should be adjusted for 0.25 VDC at the test point.

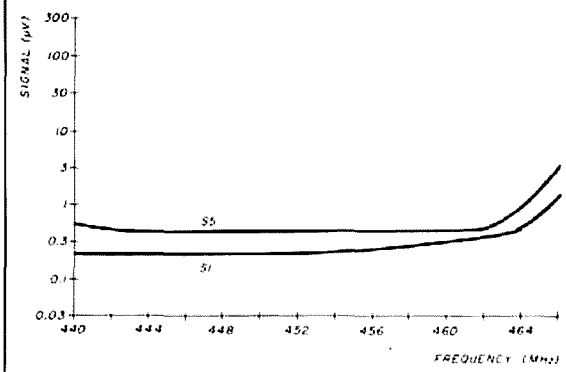


fig. 8. Graph of sensitivity vs frequency for the IC-04. Due to the higher frequencies used, the IC-04 receive bandwidth is much greater than that of the IC-02.

Now reassemble the IC-04 except for the installation of the PLL shield and back cover. Carefully peel away the metal tape covering the VCO shield. Connect a high-impedance voltmeter to the VCO test point shown in **fig. 7**. Attach battery pack, turn on the rig, and select 440.0 MHz from the keyboard.**

Adjust the capacitor through the lower right hole in the VCO shield for a reading of 0.25 VDC. This sets the VCO at its lower limit to maximize receiver cover-

*This modification is for units with serial numbers below 34,000. Units with serial numbers above 34,000 have a new CPU with different programming; diode matrix is also different. —Ed.

**If the display should power-up in the wrong 10-MHz band segment, follow the procedure under "operation" to put the rig in the proper 10 MHz segment, or press reset button on main chassis.

age above the 70-cm band. Note that this voltage is lower than the voltage for the IC-02; this is because of a smaller negative VCO shift in the IC-04 when transmitting. The IC-04 receiver will now operate from slightly below 440 MHz to slightly above 466 MHz, and the transmitter will operate cleanly throughout the upper half of the 70-cm band. (**Figure 8** shows receiver sensitivity for the entire VCO range.) Now reinstall the VCO metal tape, PLL shield, and back cover.

operation

With the modification completed, normal transmit/receive capability of the IC-02 and IC-04 is unaffected. Within a particular 10-MHz band segment, the frequency may be selected by direct keyboard entry or by using the step-up or step-down keys. To change to a different 10-MHz segment, use the step-up or step-down keys after entering the highest or lowest frequency within the present band segment. Any frequency entered will store properly in memory. If you select a receive frequency outside the VCO limits the PLL will unlock; no indication of this condition will be given on the display. However, if you push the transmit switch with an unlocked PLL, the display will blink.

conclusion

This simple modification increases the versatility of the IC-02/04 without affecting normal operation.

Incidentally, the display and microprocessor will support initialization to the 23-cm (1260-1300 MHz) band. Is this an indication of things to come?

I'll be happy to answer any questions; please enclose an SASE for a prompt reply.

ham radio

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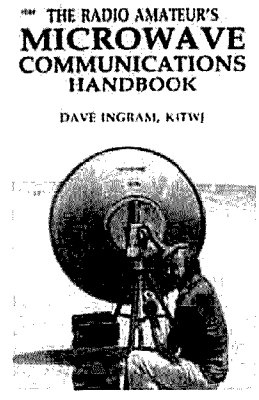
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VHF/UHF WORLD

Joe Reisert W1JR

microwave and millimeter-wave propagation: part 1

Summer is **traditionally** the time when Amateurs take to the higher bands (*and* to the hills) for mountaintopping; the weather is agreeable, and the pioneering spirit is spurred on by several major VHF/UHF contests.

Regardless of season, the frequencies above 900 MHz are still the region of exploration and discovery. This is true because, in comparison with the HF/VHF frequencies, so little Amateur work has been conducted there. Until recently, equipment was scarce and anyone interested in exploring microwave frequencies had to "roll their own" equipment.

But this is is changing rapidly. Commercial gear is now available through 13 cm (2320 MHz), as are Amateur Gunnplexers for the 3-cm (10-GHz) and 12.5-mm (24-GHz) bands.[1] One equipment manufacturer is rumored to have 3-cm weak signal gear almost ready for market!

Although much has been written on VHF and UHF propagation, information on microwave and millimeter propagation is scattered hither and yon.[2,3] With all the interest in the microwave bands, the addition of our newest band at 33 cm (902-928 MHz)[4] and the increased availability of microwave gear, this seems like a good time to start pushing these bands and promoting propagation research.

our microwave bands

Before we can talk about our microwave and millimeter bands we have to know what frequencies are available. I guess it's fair to say that — in Amateur terminology — the frequencies between 900 and 10,500 MHz are probably best referred to as the microwave bands while the Amateur frequencies above 10.5 GHz are really in the millimeter territory.

The major worldwide Amateur frequency allocations on and above 33 cm were listed in last January's column.[5] Since that time, the FCC has implemented the WARC '79 frequencies and changed the United States frequency allocations significantly. Therefore, I've updated last winter's list, (**table 1**), to show the latest USA frequency allocations. (I don't have any post-WARC '79 worldwide frequency allocation lists to compare it to at this time; I'd appreciate it if readers would share some with me.)

challenges

Once microwave and millimeter gear is constructed or purchased, there's always the question of what to do with it. Obviously it can be used in a variety of ways such as communications links, ATV, repeaters, or beacons. However, the real challenge is to see what performance can be attained primarily by DXing. This will truly exercise power, receiver sensitivity, and antenna gain to their fullest.

To show the whole perspective of the possibilities of DX on the microwave and millimeter bands, I've updated the records first shown here in last July's column.[3] **Tables 2** and **3** show the worldwide claimed DX records for terrestrial and EME operation. Note that the 13-cm band is still the upper frequency for a completed EME QSO.

In keeping with the spirit of regional DX, the North American DX records have also been updated in **table 4**. You'll note that since these records were first published, several have been broken.

Because the 33-cm band didn't even exist last summer, the DX record on that band can't begin to compare with the adjacent bands. I expect this, too, will rapidly change. Nevertheless, it offers a big challenge to see what kind of propagation is in store.

Amateur 33-cm gear is already commercially available, though it's primarily limited to the FM citizens-band type. Information on homebrew gear was recently published in this column[4]; commercial weak-signal gear is, as I indicated above, reportedly just around the corner. Because of the 33-cm band's proximity to the land mobile communications band, suitable components — especially high-power transmitting tubes — should become available; far more equipment, certainly, than can be expected to be available for the 23-cm band.

microwave and millimeter tradeoffs

These bands offer many advantages over the VHF/UHF bands. First, there's over 13,000 Megahertz — not counting the unlimited territory above 300 GHz. This is more than 275 times the bandwidth of all of the dozen lower-frequency bands combined! How long do you think it will be before microwave and millimeter wave QRM will be a problem?

These bands have very little noise

and static. Except for a small potential for VCRI around 915 MHz, TVI and RFI are virtually nonexistent. Very compact antennas are the norm, with a 1-meter diameter dish considered a moderate to large antenna, especially on 3 cm and above.

Loop Yagis are often used on the lower microwave bands. Parabolic dishes are the most common antenna type as the frequency increases because they're usually inexpensive, easy to feed, low in noise pickup, and have high gain.[6,7] As a result, QRP operation (often much less than 1 watt) can be very productive.

There are a few disadvantages, however, to operating on the microwave and millimeter wave bands. The higher antenna gain makes precise aim a serious concern. For example, a 1-meter parabolic dish antenna at 12 mm (24 GHz) has a 45-dBi gain and a beamwidth of only 1 degree!

Other problems are associated with operating above the UHF bands. Solid-state RF power generation is in the early developmental stages. Feedline losses are considerable, propagation modes are somewhat limited, and foliage attenuation and water vapor can also present difficulties.

Table 1: This table lists the latest USA frequency allocations for Amateur microwave and millimeter wave bands. Many of the listings were recently changed, as indicated in the notes.

Band	Frequencies	Notes
33 cm	902-928 MHz	Restrictions in CO, WY, White Sands, NM & Region III areas. Available as of Sept. 28, 1985.
23 cm	1240-1300 MHz	1215-1240 MHz removed on March 1, 1986.
13 cm	2300-2310, 2390-2450 MHz	2310-2390 MHz removed on November 6, 1984.
9 cm	3300-3500 MHz	
6 cm	5650-5925 MHz	
3 cm	10-10.5 GHz	
12 mm	24-24.25 GHz	
6 mm	47-47.2 GHz	48-50 GHz removed on March 1, 1986.
4 mm	76-81 GHz	71-75.5 GHz removed on March 1, 1986.
3 mm	119.98-120.02 GHz	New assignment on March 1, 1986.
2 mm	142-149 GHz	165-170 GHz removed on March 1, 1986.
1 mm	300 GHz & up	240-250 GHz removed on March 1, 1986.

Table 2: This table shows the latest claimed worldwide microwave and millimeter wave terrestrial DX records.

Frequency	Record Holder	Date of QSO	Prop. Mode	DX Miles(km)
903 MHz	AF1T-WB1FKF	1-13-86	Tropo	53(85)
1296 MHz	KH6HME-N6CA	6-24-86	Tropo duct	2472(3977)
2.3 GHz	VK5QR-VK6WG/P	2-17-78	Tropo duct	1170(1883)
3.4 GHz	VK5QR-VK6WG	1-25-86	ducting	1171(1884)
5.7 GHz	G3ZEZ-SM6HYG	7-12-83	ducting	610(981)
10 GHz	IØSNY/EA9-IØYLI/IE9	7-08-83	ducting	1032(1660)
24 GHz	I3SOY/3, IW3EHQ/3-14BER/6, I4CHY/6	4-25-83	LOS	180(289)
47 GHz	HB9AMH/P-HB9MIN/P	6-11-84	LOS	33(53)
474 THz	K6MEP-WA6EJO	6-09-79	LOS	15(24)

There is hope, however. Solid-state RF power is rapidly increasing, with power GaAs FETs (up to several watts) now in commercial production. Short feedlines are the norm, with transmitters and receivers often mounted right at or only a short distance away from the antenna feed. And many of the apparent propagation problems can be solved, as we'll see shortly.

microwave propagation

The types of propagation modes available on the microwave and millimeter wave bands are more limited than on the VHF/UHF bands. Some of the lower- frequency propagation modes such as meteor scatter, sporadic E, F2, TE, and FAI, for example, are not usable.[2,3] Other exciting scatter propagation modes are available, however, that may be either poor or simply not feasible on the VHF/UHF bands.

The most common microwave propagation modes are line-of-sight (LOS), tropospheric bending, tropospheric ducting, and EME. Less used but readily accessible are various scatter modes using the troposphere, weather-related phenomena such as lightning and rain scatter, aurora, and both man-made and natural objects.

Because describing the various microwave and millimeter wave propagation modes and how they can best be exploited would take considerable space, only LOS microwave and millimeter-wave propagation will be discussed this month. The other modes will be discussed in detail in next month's column.

line-of-sight propagation

The most common mode of propagation on the microwave and millimeter wave frequencies is probably LOS, for apparently obvious reasons. If two stations' antennas can see each other without intermediate obstructions, it is assumed that communication is possible.

This does, however, assume several things. First, there must be sufficient transmitter power and receiver sensitivity for communication to take place.

Second, the gain of the antenna at both ends of the path must be sufficient. And third, the weather must be cooperative. (More on this next month.)

But what is sufficient power, sensitivity, and antenna gain? These parameters are all interrelated. If sufficient RF power — say 100 to 1500 watts — is available, the effective radiated power, even from a low-gain antenna (less than 10 dBi) will be high.

Conversely, if transmitter power is low, even in the milliwatt range, communication is also possible, providing that the gain of the antennas at both ends of the path is sufficient and that the receiver has sufficient sensitivity. These tradeoffs will be addressed in a future column. This month's column will be limited to a discussion of "path loss," the apparent number of dBs of attenuation that a transmitted signal will incur between two isotropic antennas. This is the standard method used in the communications industry to determine whether there will be a high probability that communciations is possible between two stations.

Table 3: This table shows the latest claimed worldwide microwave EME DX records.

Frequency	Record Holder	Date of QSO	Prop. Mode	DX Miles(km)
903 MHz	None reported.			
1296 MHz	PAØSSB-ZL3AAD	6-13-83	EME	11595(18656)
2304 MHz	PAØSSB-W6YFK	4-05-81	EME	5491 (8836)
3300 MHz and above	none reported.			

Table 4: This table shows the latest claimed North American microwave and millimeter wave DX records listed by suspected propagation modes. Note that most of these records are far short of the worldwide claims and therefore offer a great challenge to North American Amateurs. (Note: the records are listed alphabetically by propagation mode. Where the path is mostly over water, ducting is suspected. No efforts are made to separate out ducting on overland paths, so it's grouped under tropo.)

Frequency	Record Holder	Date	Prop.Mode	DX Miles(km)
903 MHz	AF1T.WB1FKF	1-13-86	Tropo	53 (85)
1296 MHz	KH6HME-N6CA	6-24-84	Ducting	2472 (3977)
	K2UYH-VK5MC	12-06-81	EME	10,562(16,995)
	W4WSR-WA5TKU	6-03-85	Tropo	1112 (1790)
2.3 GHz	PAØSSB-W6YFK	4-05-81	EME	5491 (8836)
	W4ODW-WB5LUA	2-20-86	Tropo	624 (1003)
3.4 GHz	K6HIJ/6-W6IFE/6	6-18-70	LOS	214 (344)
5.6 GHz	K5FUD-K5PJR	9-20-77	Tropo	267 (430)
10 GHz	WA4GHK/4-WD4NGG	8-07-84	Ducting	297 (478)
	W7JIP/7-W7LHL/7	7-31-60	LOS	265 (426)
24 GHz	KXØO/Ø, WØMXY/Ø-NKØP/Ø, WAØVSL/Ø	8-24-85	LOS	74 (119)
48 GHz	W2SZ/1-WA2AAU/1	9-08-84	LOS	0.3 (0.5)
50-300 GHz	None reported.			
474 THz	K6MEP-WA6EJO	6-09-79	LOS	15 (24)

path loss

What is path loss and how is it determined? We all know that as you go higher in frequency, losses become more critical. We also hear that the path losses increase. Yet at the same time, we hear about tremendous DX accomplishments on the microwave frequencies, using low power and relatively small antennas. How can this be?

The main reason for this is the "wavelength factor." As you go higher in frequency, the size of a half-wavelength dipole antenna becomes smaller. Therefore its "capture area" becomes smaller[8] and it consequently picks up less RF. For example, the "effective aperture" of a half-wavelength dipole is approximately 0.75 by 0.25 wavelengths.[8] At 33 cm this represents a capture area of about 32 square inches (0.21 square meters). However, at 3 cm (10 GHz) the effective aperture is still 0.1875 square wavelengths,[8] but because of the shorter wavelength, the capture area is reduced to 0.2612 square inches (0.00017 square meters), a reduction of over 120 times.

It's said that the answer to this dilemma is simple: just increase the capture area of the antenna commensurately and you'll gain back what you lost because of the wavelength factor. This is only partially true. Antenna gain is usually a two-way street, receiving and transmitting (if the same antenna is used for both). Hence, if the antenna size is increased to offset the capture area loss, you gain considerable advantage, depending on frequency.

Herein lies the secret of successful use of low power on the microwave and millimeter frequencies. If the antenna gains is increased sufficiently, the overall signal-to-noise ratio can actually increase for the same transmitter power and receiver sensitivity.

How is the path loss determined? There are straightforward formulas for this. The most common one is:

$$Path\ loss = 37.6 + 20 \log F + 20 \log R \quad \text{(eq 1)}$$

where path loss is in dB referenced to an isotropic radiator (2.15 dB less gain than a dipole), F is frequency in MHz, and R is range in miles.

If you prefer to work in kilometers, the formula can be modified as follows:

$$Path\ loss = 33.6 + 20 \log F + 20 \log R \quad \text{(eq 2)}$$

where path loss is in dB referenced to an isotropic radiator, F is frequency in MHz, and R is range in kilometers.

For example, the path loss for 10 miles is approximately 116.7 dB at 903 MHz and 137.6 dB at 10 GHz. If the range is increased to 20 miles, the path loss is 122.7 and 143.6 dB, respectively. You'll note that every time the fre-

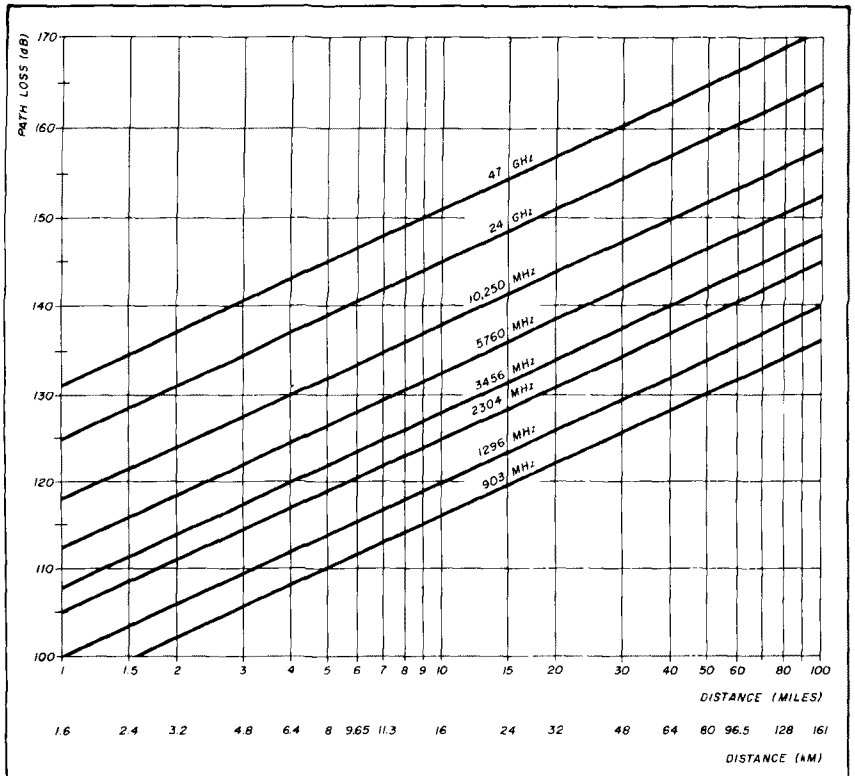


fig. 1. This graph was derived using equations 1 and 2. It shows path loss versus frequency of operation and range between two stations. This assumes that there is a line-of-sight path and no obstructions or weather phenomena are present.

quency or range is doubled, the path loss increases by 6 dB.

Calculations of path loss are often computer-aided.[9] For those not so inclined, I've prepared **fig. 1**, which shows at a glance the path loss on each of the microwave and lower millimeter Amateur bands.

summary

This month's column served as an introduction to radio propagation on the microwave and millimeter wave Amateur bands. FCC frequency allocations and DX records were also shown. The latter will quickly reveal the challenges available to the adventuresome Amateur.

Time and space allowed us to investigate only line-of-sight propagation, presently the most common use of these bands. Next month's column will explore the other modes of propagation, with particular emphasis on those useful for DXing. A review of references 2 and 3 is strongly recommended.

references

1. Joe Reisert, W1JR, "VHF/UHF World: VHF/UHF Exciters," *ham radio*, April, 1984, page 84.
2. Joe Reisert, W1JR, "VHF/UHF World: The VHF/UHF Primer — An Introduction to Propagation," *ham radio*, July, 1984, page 14.
3. Joe Reisert, W1JR, "VHF/UHF World: Propagation Update," *ham radio*, July, 1985, page 86.
4. Joe Reisert, W1JR, "VHF/UHF World: 33 cm — Our Newest Band," *ham radio*, April, 1986, page 83.
5. Joe Reisert, W1JR, "VHF/UHF World: Amateur Microwave Bands," *ham radio*, January, 1986, page 44.
6. Joe Reisert, W1JR, "VHF/UHF World: Reflector Antennas — Part 1," *ham radio*, February, 1986, page 51.
7. Joe Reisert, W1JR, "VHF/UHF World: Reflector Antennas — Part 2," *ham radio*, March, 1986, page 68.

Important VHF/UHF Events

July 2: *±2 weeks. Look for European openings on 6 meters.*
July 18-20: *CQ VHF WPX contest.*
July 19: *EME perigee*
July 20: *Look for 2-meter sporadic E ±2 weeks of this date.*
July 25-27: *Central States VHF Conference, St. Louis, MO (contact NØIS).*
July 29: *Predicted peak of the Delta Aquarids meteor shower at 0930 UTC.*
August 2-3: *ARRL UHF Contest.*
August 12: *Predicted peak of the Perseids meteor shower at 0700 UTC.*
August 16: *EME perigee*
August 28-31: *Microwave Update 1986 Conference, Estes Park, CO (contact WØPW).*

ham radio

short circuits

meteor shower program

In a footnote to W1JR's June column, "VHF/UHF World: Meteor Scatter Communications" (page 68), the address of Gary Field, WA1GRC, was shown incorrectly. Gary's correct address is 5 Pluff Avenue, North Reading, Massachusetts 01864.

solving transmission line problems on the C-64

Gary Myers, K9CZB, author of "Solving Transmission Line Problems on the Commodore 64" (May, 1986, page 74), recently learned that some of the older C64s have an operating system that doesn't poke the graphic symbol in lines **180-240** properly. On these machines, the arrow symbol is generated in the same color as the background, thus rendering it invisible. This won't affect the program's operation, but there will be no visual confirmation that the desired key was pressed.

Gary reports that WB9PGO has devised a program modification that will make the symbol visible on these machines. Those who don't see the graphic arrow after selecting the transmission line type should change line **20** to read as follows:

```
20
POKE53281,14:PRINTCHR$(147):
POKE53281,6
```

and change the last statement in line **670**, which currently reads **GOTO70**, to **GOTO20**.

credit where it's due

Credit lines were inadvertently omitted from the art in Walter Kunde's June article, "Direct Currents Reduce Core Permeability" (page 58). All figures appeared courtesy of Magnetics, Inc., P.O. Box 391, Butler, Pennsylvania 16003.

a tone burst generator for European repeaters

Use 7168 kHz crystals, divided down output for 1750 Hz output

Before a recent trip to Europe, I received licenses to operate and decided to take my ICOM 2AT along with me. While searching through *Radio Communication* and other journals to find information on repeaters, I noticed that most European repeaters must be brought up with a 1750 Hz tone. Obviously my 2AT could not do this.

Looking through the ads, I noticed that ICOM makes a European version of the 2AT called the ICOM 2E. Besides offering coverage from 144 to 145 MHz only, this model appeared to employ either a switch-operated 1750 Hz tone or a 1750 burst at the beginning of each transmission. A phone call to ICOM confirmed this. I considered converting my 2AT to a 2E . . . but the touchtone pad on the 2AT is located where the burst components would have to be and uses several tracks on the circuit board for different functions. In addition, the tone switch on the 2AT is a push-button switch activated by pushing the volume control; one look into the possibility of removing that control eliminated that idea. I concluded that while conversion would be possible, it wouldn't be reasonable to make the conversion, use the unit abroad, and then reconvert it when I returned home.

My initial thought was to use a 555 timer. But I wasn't sure it would survive the trip and still make the 25 Hz frequency tolerance required. Without test equipment, I would be unable to make any repairs while in Europe.

The ICOM service manual for the 2AT showed that the tone burst circutry consisted of a single integrated circuit, a crystal, and several discrete components. The integrated circuit, which functions as an oscillator/divider, was a TC5082 packaged in a 9-pin single-in-line package with an SK3733 indicated as its replacement. A quick phone call showed the device, also known as an ECG1197, was available locally for under $10.00. It contains a crystal oscillator and divide-by-256, 1024, 2048, and 4096 and is often used in CB radios as a reference oscillator for the PLL.

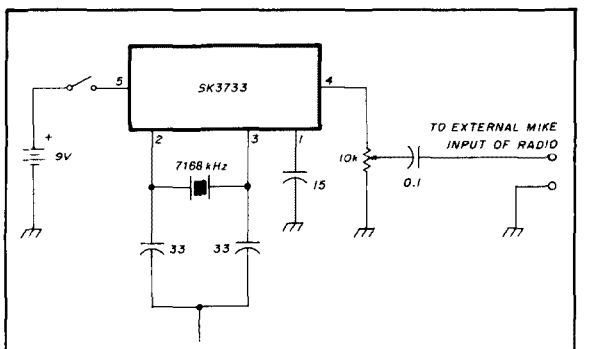


fig. 1. Tone generator produces 1750 Hz output starting with 7168 kHz crystal frequency.

The final circuit, shown in **fig. 1,** is almost identical to the ICOM burst generator. A 7168-kHz crystal was used so the divide-by-4096 output would be 1750 Hz. The output was capacitively coupled to the external microphone jack of the 2AT. Since the 2AT requries DC continuity for transmit, the PTT bar on the side had to be pushed for transmit.

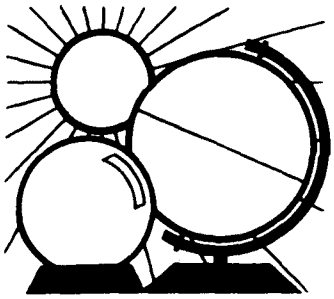
Construction was on a small piece of 3-pad-per-hole board. The chip is CMOS, so standard CMOS handling procedures — i.e., using a grounded soldering iron, inserting the chip last, etc. — were used. The case, recycled from an old GE pager, had enough room for the circuit and a 9-volt battery. The press-to-silence switch already on the case was used to activate the burst. A short audio cable coupled the two units.

Tuneup was easy. I just asked a friend to listen to my simplex signal and report on its loudness. I adjusted the pot until the level was approximately the same as my voice or the touchtone pad.

Coupling to another rig should be simple. Since most HTs have external microphone jacks, that should be the way to go. One could build an adapter to enable this unit and the external microphone to be used simultaneously.

ham radio

By Ladimer S. Nagurney, WA3EEC, 73 Blackberry Lane, Amherst, Massachusetts 01002



DX FORECASTER

Garth Stonehocker, KØRYW

a geomagnetic-ionospheric disturbance

Because the full effects of solar flares and storms will be upon us again in a year or so, a brief review of the solar flare-to-ionospheric storm sequence may be beneficial, particularly for those who've joined the DXing ranks since 1982.

Several years have passed since the sun has been active enough to cause concern over the weak signals resulting from geomagnetic-ionosopheric disturbances in any great number or intensity. Although one or two disturbances a year have been of the geomagnetic-ionospheric variety, most of those we've experienced during the last four years have been mild solar wind enhancements. However, single events — such as that witnessed from the 5th through the 10th of February, 1986 — do occur unexpectedly.

When a sunspot region flares or brightens, ultraviolet and x-rays cause increased ionization in the D and E regions of the ionosphere. Signals are immediately weakened on the daylight side of the earth. This sudden ionospheric disturbance lasts about an hour. With the eruption of the flare, many solar particles (protons and electrons) ejected into the solar wind start toward the earth along a spiral path toward the polar regions. The higher-energy protons arrive within approximately one to five hours, causing polar cap absorption — even blackout — of signals crossing the top and bottom of the earth above 80 degrees latitude during each period of daylight for a couple of days. The more numerous, but slower, electrons arrive within 20 to 30 hours; their arrival often causes a shock wave which suddenly begins to move the geomagnetic field around, then lasts for two to three days. This phenomenon is called SC, or "sudden commencement." The presence of many particles results in further weakening and level fluctuation of signals as a result of movement in the auroral zone (70 to 80 degrees latitude).

The ionospheric F region is affected after the SC, first by an increase in maximum usable frequency (MUF) for a few hours over the transmission path that's in sunlight. Over the path that's in darkness, the MUF decreases from the auroral latitudes southward, depending upon the intensity of the storm. (See the MUF distribution tables in the January and February, 1986, columns.)

Figure 1 provides a graphic representation of the effects of the February 1986 ionospheric disturbance. The top curve (**fig.1A**) is the percentage change in foF2; the middle curve (**fig. 1B**) is the solar flux value; the lower curve (**fig. 1C**) is the geomagnetic A index for each day during February. By looking at each day individually and then reading across the days, one can see the storm variation. To see how the storm actually developed from February 5th to the 10th, see **table 1**, which shows the percentage of foF2-MUF change, and in parenthesis the geomagnetic K figure associated with that 2-hour period. The median foF2 for that hour of the month is indicated in the bottom line of the table.

Table 1. Percentage of MUF change and geomagnetic K figure (in parentheses) associated with 2-hour period.

Date	1300 UT	1500	1700	1900	2100
5	5.9(1)	0.0(1)	6.3(2)	0.0(2)	0.0(2)
6	9.8(1)SC 1313	5.3(3)	7.9(3)	22.6(2)	5.1(3)
7 SC 0500	3.9(5)	-8.8(5)	7.9(6)	11.4(6)	-11.9(7)
8	-19.6(7)	-38.6(8)	-20.6(8)	-32.3(9)	-25.7(9)
9	-19.6(4)	-5.3(4)	-15.9(5)	-6.5(5)	3.4(3)
10	-11.8(1)	-8.8(1)	-14.3(1)	-8.1(3)	-5.1(3)
Median foF2	5.1	5.7	6.3	6.2	5.9

The storm MUF decrease was probably not as extreme as the geomagnetic field A and K figures indicate because the solar flux was high and the initial SC 1313 UT was coincident with sunrise, the foF2 rise of the day. A second SC was on the 7th at about 0500 UT; the main decrease in foF2-MUF began about 1100 UT with the large K figures. The 8th, which was a Saturday, was particularly bad, as anyone trying to communicate over any significant distance will remember. Several days later MUFs returned to normal. Such is the morphology of a geomagnetic-ionospheric disturbance.

last-minute forecast

The higher frequency bands, 10 through 20 meters, are expected to improve during the first week of the month. They should be best during the

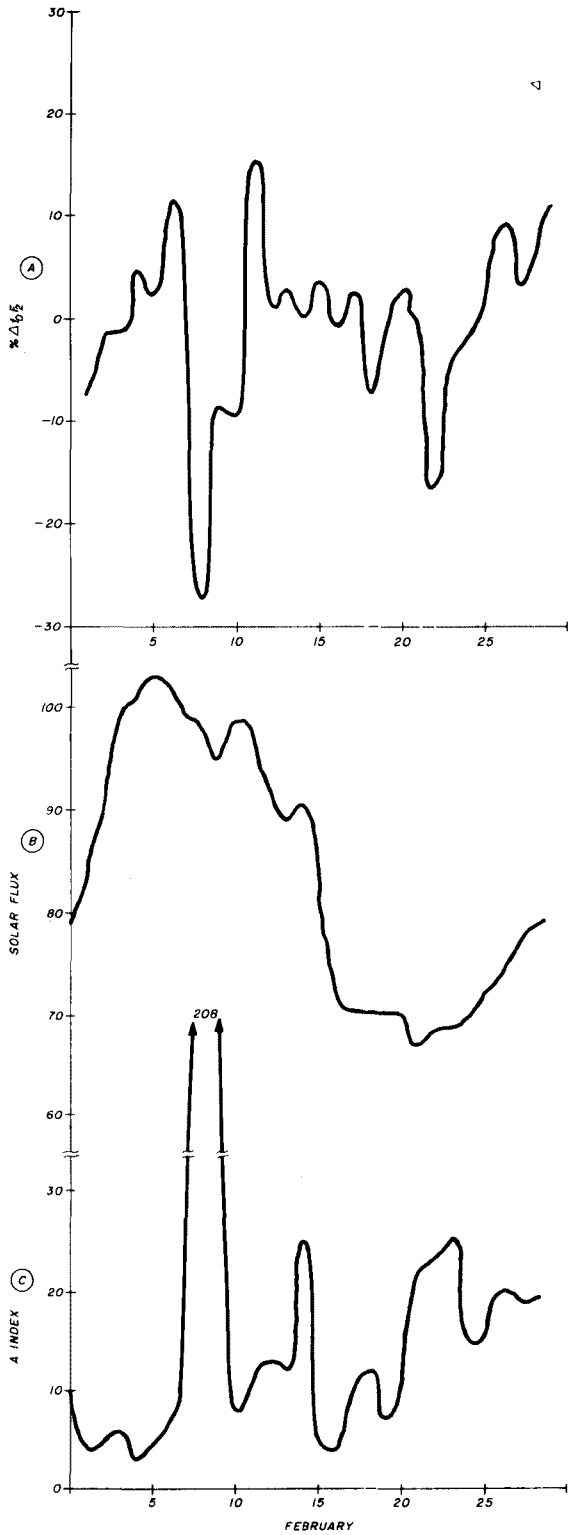


fig. 1. Pronounced effect of solar flare on: (A) MUF (percentage change in MUF); (B) solar flux; and (C) geomagnetic field "A" index.

second and third weeks, when the solar flux 27-day maximum is expected. Though still not very good, the lower bands will be best during the fourth and last weeks of the month. Disturbances are more probable on the 3rd to the 4th, the 17th to the 18th, and on the 23rd through the 25th during the solar flux transition and minimum time periods. Lunar perigee (closest approach of the moon) is on the 19th, with a full moon occurring on the 28th. The Aquarids meteor shower begins July 18, peaks on the 28th, and lasts until August 7. The radio-echo rate at maximum is expected to be about 34 per hour.

band-by-band summary

Six meters will have occasional openings to South Africa and South America around local noon via E_s short skip. These can occur at any time of the month and favor a five-to-six day cycle.

Ten and fifteen meters will have many short-skip openings near local noontime and long-skip, especially during any solar flux rise, to most southern areas of the world during daylight. No long-hop transequatorial openings are expected to occur at this time of the year.

Twenty, thirty, and forty meters will have DX conditions from most areas of the world during the daytime and into the evening almost every day, either long-skip to 2500 miles per hop or short-skip E_s to 1000-mile hops. The length of daylight is still near maximum, providing many hours of high maximum usable frequencies for good DXing.

Thirty, forty, eighty and one-sixty meters are all good for nighttime DX if you can beat the buildup of thunderstorm QRN in the evenings. Some enhanced signal strength levels via short-skip E_s may help overcome the noise. Some operators get up during the predawn hours, after the thunderstorms have dissipated, taking advantage of quieter conditions to the east. Time-and-frequency station MSF, in Rugby, England, can be used to monitor 2.5 and 5 MHz for band conditions.

JULY

GMT	PDT	WESTERN USA N	NE	E	SE	S	SW	W	NW	MDT	MID USA N	NE	E	SE	S	SW	W	NW	CDT	EDT	EASTERN USA N	NE	E	SE	S	SW	W	NW
0000	5:00	20	30	20	12	20	10	10	15	6:00	20	30	20	12	20	10	10	15	7:00	8:00	20	20	20	12	20	10	10	15
0100	6:00	20	*30*	20	12	20	10	10	15	7:00	20	30	20	12	20	10	10	15	8:00	9:00	20	20	20	12	20	10	10	15
0200	7:00	15	*30*	*20*	15	20	10	10	15	8:00	15	*30*	*20*	15	20	10	10	15	9:00	10:00	20	20	*20*	15	20	10	10	20
0300	8:00	15	*30*	*30**	20	20	10	10	15	9:00	15	*30*	*30**	20	30	10	10	20	10:00	11:00	20	20	*30**	20	30	10	10	20
0400	9:00	20	*30*	*30*	20	30	10	10	15	10:00	20	*30*	*30*	20	30	10	10	20	11:00	12:00	20	*20*	*30*	20	30	12	12*	20
0500	10:00	20	20	*20*	20	30	10	10	15	11:00	20	*30*	*20*	20	30	12	12	20	12:00	1:00	*30*	*30*	*20*	20	40	12	12	20
0600	11:00	20	20	15	20	30	12	12	20	12:00	*20*	20	20	20	40	12	12	20	1:00	2:00	*30*	*30*	20	20	40	15	15	*20*
0700	12:00	20	20	15	*20*	30	12	12	20	1:00	*30*	20	20	*20*	40	15	15	20	2:00	3:00	*30*	*30*	20	*20*	40	20*	15	*20*
0800	1:00	20	20	20	*20*	40	15	20	20	2:00	*30*	20	20	*20*	40	20	20	*20*	3:00	4:00	20	20	20	*20*	*40*	20	20	*30**
0900	2:00	*20*	20	20	*30**	40	15	20	20	3:00	*20*	20	15	*30**	*40*	20	20	*30**	4:00	5:00	20	20	20	*30**	*40*	20	20	*20*
1000	3:00	*30*	20	20	*30*	*40*	20	20	20	4:00	20	20	15	*30*	*40*	20	20	*20*	5:00	6:00	20	20	15	20	*40*	20	20	20
1100	4:00	*30*	20	20	20	*40*	20	20	*30*	5:00	20	20	12	20	*40*	20	20	*20*	6:00	7:00	20	20	15	20	*40*	20	20	20
1200	5:00	*20*	20	15	20	*40*	*20*	20	*30*	6:00	20	20	12	20	*40*	*20*	20	20	7:00	8:00	20	20	12	15	*40*	20	20	20
1300	6:00	20	20	12	20	*40*	*20*	20	*30*	7:00	20	20	10	15	*40*	*20*	30	20	8:00	9:00	15	20	12	15	*40*	*20*	30	20
1400	7:00	20	20	12	15	*40*	*20*	*30*	20	8:00	20	20	10	15	*40*	*20*	*30*	20	9:00	10:00	20	20	10	12	*40*	*20*	*30*	20
1500	8:00	20	15	10	15	*40*	*20*	*30*	*30*	9:00	20	20	10	12	*40*	*20*	*30*	*20*	10:00	11:00	20	20	10	12	*40*	*20*	*30*	*20*
1600	9:00	20	15	10	12	*40*	*20*	*30*	*30*	10:00	20	15	10	12	*40*	*20*	*30*	*20*	11:00	12:00	20	20*	10	10	*40*	*20*	*30*	*20*
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1800	11:00	*20*	15	12	10	20	15	*30*	20	12:00	20	15	12	10	20	15	*30*	20	1:00	2:00	20	15	12	10	20	15	*30*	20
1800	12:00	*30*	20*	12	10	20	12	20	20	1:00	*20*	20	12	10	20	12	20	20	2:00	3:00	*20*	20	12	10	20	12	15	20
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2100	2:00	20	20	20	10	20	10	10	20	3:00	*30*	20	20	10	20	10	10	20	4:00	5:00	*30*	20	20	10	20	10	10	20
2200	3:00	20	20	20	10	20*	10	10	20	4:00	20	20	20	10	20	10	10	20	5:00	6:00	20	20	20	10	20	10	10	20
2300	4:00	20	20	20	10	20*	10	10	20	5:00	20	20	20	10	20	10	10	15	6:00	7:00	20	20	20	10	20	10	10	15
		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN			ASIA FAR EAST	EUROPE	S. AFRICA	CARIBBEAN S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN

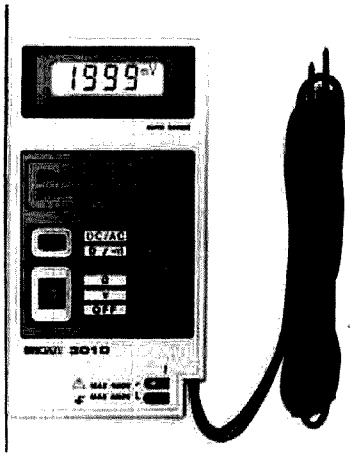
The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.

*Look at next higher band for possible openings.

NEW products

personal pocket DMM

This compact, inexpensive, new instrument from North American Soar features simplicity of operation and rugged design. The fully autoranging Model 3010 doesn't require fuses because it's electronically protected from operator misuse. Even applying 250 volts AC to this meter in the ohms position won't hurt it.



Several features, such as permanently mounted test probes that can't be lost or connected incorrectly, have been added to ensure good performance in the field. The lightweight design, durable ABS plastic housing, and the carrying case all afford "drop-proof" protection at modest cost: only $29.95.

For further information, contact North American Soar Corporation, 1126 Cornell Avenue, Cherry Hill, New Jersey 08002.

Circle #302 on Reader Service Card.

new IC-28A and IC-28H 2-meter mobiles

ICOM has announced the availability of the new IC-28A 25-watt and IC-28H 45-watt packet-compatible 2-meter rigs with all the features necessary for mobile operation. These features include compact size (the IC-28A measures 5-1/4 x 5-1/2 x 2 inches, the IC-28H 7-1/4 x 5 1/2 x 2 inches) and a large LCD readout with an automatic dimmer circuit to reduce brightness.

The units operate from 138-174 MHz, with specifications guaranteed from 144.00-148 MHz, making them ideal for MARS and CAP operation. Twenty-one memory channels are included. It's possible to scan the entire band or the memory channels from the provided HM-12 mic. Easy to operate, each unit features only 11 front panel controls.

Options include the IC-HM14 DTMF mic, PS-45 13.8V, 8A power supply, UT-29 tone squelch unit, SP-10 external speaker, HM-16 speaker mic and HS-15/HS-15SB flexible boom mic, and PTT switchbox.

For information, contact ICOM America, Inc., 2380 116th Avenue N.E., Bellevue, Washington 98009-9029.

Circle #303 on Reader Service Card.

antenna switch

MFJ Enterprises, Inc., is now producing the MFJ-1701, a six-position antenna switch that allows switching antennas with the turn of a knob. It organizes tangles of coax cables and eliminates the need to keep plugging in and unplugging cables.

The MFJ-1701 retails at an affordable $29.95 (plus $5.00 shipping and handling). The equipment is mounted in a rugged, yet handsome, black aluminum cabinet that matches most rigs.

This six-position antenna switch has SO-239 connectors, negligible insertion loss, low VSWR, and low crosstalk between adjacent outlets. All unused terminals are automatically grounded for static/lightning/RF protection. The MFJ-1701 can be used for 52- to 75-ohm systems and can be mounted with equal ease on a desk or on a wall. In addition, the MFJ-1701 handles 2000 watts SSB or 1000 watts CW.

This product is backed by MFJ's one-year unconditional warranty. If ordered directly from MFJ, it has an additional 30-day guarantee — return it within 30 days for a full refund (minus shipping and handling) if not completely satisfied.

For details, contact MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, Mississippi 39762.

Circle #304 on Reader Service Card.

new amateur equipment from Heath

Four new Amateur Radio kit products have been introduced by Heath Company, the world's largest manufacturers of high-technology electronic kit products. The new products are the HD-1420 VLF Converter, HD-1422 Antenna Noise Bridge, HD-1424 Active SWL Antenna, and the HD-1530 Touch-Tone Decoder.

The HD-1420 Very Low Frequency (VLF) Converter allows a standard shortwave receiver to tune the 10 to 500 kHz band using the receiver's 3.5 to 4.0 MHz band.

The HD-1422 Antenna Noise Bridge is a useful antenna tuning aid which reveals the cause of any mismatch between a station's transmitter and its antenna.

The HD-1424 Active SWL Antenna allows a shortwave radio to receive signals between 300 kHz to 30 MHz.

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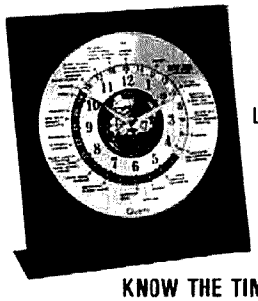
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HF2V	80-40 vertical	114.00
2MCV5	2MT vertical	52.95
RMKII	roof mtg kit	41.95
TBR160S	160m add on	45.00
MPS	mtg. post sleeve	5.50
	AND MORE!	

CUSHCRAFT

A4	4 el. triband	300.00
A3	3 el. triband	224.00
R3	10,15,20 remote tuned vert	275.95
AV5	5 band trap vert	105.00
32-19	19 el. 2mt. boomer	96.95
215WB	15 el. wide band 2mt boomer	79.95
424B	24 el. 70cm boomer	82.95
416TB	16 el. OSCAR 435 MHz	60.00
A144-10T	10 el. OSCAR 145.9 MHz	53.00
AOP-1	OSCAR pack 2mt & 70cm	150.00
AR-2	2mt. vert. ringo	24.50
ARX-2	2mt. vert. ringo ranger	30.00
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RM75/RM80	75 or 80 std	18.95
RM75S/RM80S	75 or 80 super	36.95
BM-1	bumper mt.	15.95
SSM-2	stainless ball mt	17.95
SSM-1	stainless ball & spring mt	32.95
QD-1	quick disconnect	13.95
SGM-2	2mt. 5/8 mag. mt	28.95
HOT	trunk mt. w/swivel ball	16.95
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TH5MK2S	5 el. triband	457.95
EX-14	4 el. triband	362.95
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18AVT/WBS	5 band trap vert	121.95
14AVQ/WBS	4 band trap vert	74.95
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HB144MAG	2mt mag mt.	21.95
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KT34XA	triband 5 el	485.95
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2M-22C	2mt. satellite	117.55
435-18C	70cm satellite	113.95
435-40CX	70cm satellite	155.95

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CL33	3 el. triband	265.00
PRO37	7 el. triband	465.00

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SQY-06	2mt. Swiss Quad	55.95

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Radios

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R71A HP XF 8 pole Xtal Filter add $250.00 to R71A.
R71A HP XFS 2.1 kHz Xtal Filter add $300.00 to R71A.
R71A 24 hour tested—no mods $669.95

ICOM R7000 VHF/UHF $849.00

EEB will offer a R7000 HP—Detail Release date July 86.

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NEW products

The HD-1530 Programmable Touch-Tone Decoder is used in series with the speaker of a receiver or scanner.

These Amateur Radio products are just a few of many products offered by the Heath Company. To receive a free copy of the current catalog, contact Heath Company, Department 150-592, Benton Harbor, Michigan 49022. (In Canada, write Heath Company, 1020 Islington Avenue, Department 3100, Toronto, Ontario, M8Z1).

Circle #305 on Reader Service Card.

miniature VHF GaAs FET preamps

Low in cost and small in size, these new amplifiers from Hamtronics are easily used in a variety of applications. Amplifiers in the LNW series, available in eight different models tuning from 25 to 500 MHz, are designed around dual-gate FETs that provide stable operation over a wide range of load impedances. At frequencies below 120 MHz, gain runs from 22 to 28 dB with a typical noise figure of 0.8 dB.

The price is $19 (in kit form) and $34 (assembled and tested). Further details on these preamps and many other Hamtronics products are available in a 40-page catalog available for $1. Contact Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535.

new software for RC-850 repeater controller

Advanced Computer Controls, Inc. has announced Version 3.4 software that adds many new features to repeater systems.

The controller's autopatch is enhanced to support multiple telephone lines, including up to three remote phone lines linked by radio. Remote phone lines allow autopatch and autodial services on repeaters at inaccessible sites and allow the repeater's patch coverage to match its RF coverage, with calls automatically directed to the proper line.

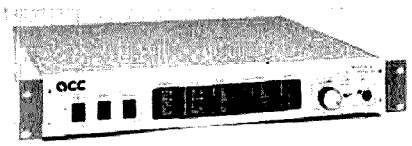
Additional access and control modes allow custom tailoring of PL and touchtone access to the repeater. Eight hundred individual user-access codes may be enabled and disabled by the repeater owner, for secure controlled access to selected functions.

Four links or remote bases are supported, with touchtone command entry permitted from the remotes. Additional remote base frequency memories are included, with programmable memory name responses.

Numerous additional paging formats are supported for selective call to users with decoders or paging receivers. New formats include five-

tone sequential, DTMF, CTCSS, and HSC for numeric display paging. Pager information is stored in memories for easy two-digit reference to any pager.

Available to all current owners, the software is supplied in 48K of EPROM, plus an 8K RAM upgrade. The non-volatile EPROM memory is optionally upgradable to 8K. The upgrades are included as standard in newly shipped equipment.



For additional information, contact Advanced Computer Controls, Inc., 10816 Northridge Square, Cupertino, California 95014.

Circle #306 on Reader Service Card.

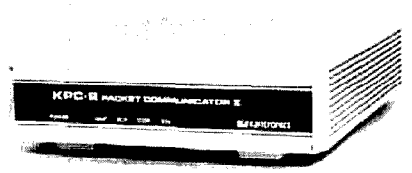
Kantronics Packet Communicator II™

The complete Kantronics Packet Communicator II (KPC-2) is an AX.25 Version 2.0 TNC that features a completely new design, the latest in technology updates, and over 100 software commands. A serial RS232 or TTL port allows connection to any computer. KPC-2 is also compatible with existing TNCs. The unit is ready to use and easy to operate.

Priced at $219, the KPC-2 is compatible with any computer having a serial, asynchronous I/O. It offers standard computer-to-communicator baud rates between 300 and 9600, and packet radio baud rates of 300, 400, 600, and 1200.

Software selectable VHF and HF modems are included. Six software selectable tone pairs including Bell 103, 202, CCITT V.21, and CCITT V.23, and full duplex capability are also featured. Full 16K is RAM standard, with memory expandable to 32K RAM.

Any terminal or communications software program can be used to establish communication between your computer and the KPC-2.™ Kantronics offers *Pacterm*™, a special packet terminal program for many popular computers.



Power supply, connectors, and cables are provided with the KPC-2, but the user must supply the transceiver mic jack and the computer RS232/TTL connector.

For information contact Kantronics, Inc., 1202 E. 23rd Street, Lawrence, Kansas 66046.

Circle #307 on Reader Service Card.

new Transi-trap™

Alpha Delta has just announced availability of a new improved vesion of its Transi-trap™ electrical surge protector.

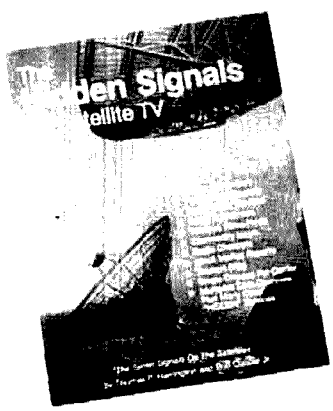
The Transi-trap "Arc-plug" has been redesigned to meet industrial and governmental protection standards for protection against Electromagnetic Pulse (EMP), in accordance with the National Communications System report, NCS TIB 85-10. The new "Arc-plug" has a DC clamping level of 350 volts to provide proper transmitter protection. The pulse clamping level (per NCS EMP test: 4,500 volts at 50 ohms) is 230 volts. The unit will respond in 80 to 100 nanoseconds and has a very low interelectrode capacitance of less than 1 pF.

For more information, contact Alpha Delta, P.O. Box 571, Centerville, Ohio 45459.

Circle #308 on Reader Service Card.

hidden signals

Universal Electronics has announced the release of the second edition of *The Hidden Signals on Satellite TV*, the first book to completely cover the field of non-video satellite services carried on domestic satellites.



These services include stereo subcarriers, telephone channels, world news and press services, Teletext, and other VBI Systems, Single Channel Per Carrier (SCPC) Systems, plus other data and business services.

Hidden Signals deals with all phases of this expanding side of the satellite business: the systems, how they work, who uses them, how they are received, and how the services can be utilized. Despite its sophisticated content, the illustrated 240-page book is easy to read and understand.

The book is available for $19.95, plus $2.00 shipping and handling from Universal Electronics, Inc., 4555 Groves Road, Suite 13A, Columbus, Ohio 43232. (Also available from Ham Radio's Bookstore, Greenville, New Hampshire 03048 for $19.95 plus $3.50 shipping and handling.)

Circle #309 on Reader Service Card.

ICOM REPEATER PROGRAM UPGRADE

The RPS-1 is an entirely new repeater program written for the ICOM REPEATER controller board that will add NEW FEATURES and CUSTOM PROGRAMMING to your ICOM RP-1210, RP-2010, and RP-3010. The RPS-1 will take your generic sounding repeater and give it a whole new personality.

New CW ID operation provides two separate cw id messages, and can include city/state, or club name. Plus the cw messages are sent at the speed you request. The new Courtesy Beep tone prompt allows a short courtesy wait period before beeping and resetting the timeout timer — while the repeater transmitter stays on. New timeout operation clears timeout immediately with no down delay, and provides cw warning messages to reduce confusion. Custom programming of all cw and timing parameters.

The ICOM REPEATER is a well built, reliable repeater, and adding the PROCESSOR CONCEPTS RPS-1 will add new features and personalized programming to make a good repeater even better.

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ATV

This converter covers all the 420 - 445 MHz ATV frequencies. This converter uses a crystal reference to lock onto the the 439.25 MHz standard - or - a second channel at 421.25 or 426.25 or 436.25 MHz. This converter also supports a manual tuning mode for non-standard frequencies. The IF output can be ordered for channel 2 or 3. This ATV module is small and only requires +12VDC. A small 12V transformer is supplied with this assembled and tested converter. This converter uses a GAS FET RF amp with a highly selective bandpass filter that will cut out strong UHF TV and other strong out-of band RFI.

Typical rejection @ 500 MHz = -50 db, @220 MHz = -65 db, and @144 MHz = -70 db or greater. The pass-band is 420-445 MHz.

Voltage regulation is very stable under varying DC input. Conversion gain is approximately 20 db with input and output surge protection. Input is via type "N" connector. Output is via a type "F" connector. Size is 5.125" x 5.5" x 1.5".

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2 " "	80, 40M	85 ft "	$ 55 "
9 BAND SPACE-SAVER DIPOLE *	160 thru 10M *	401 ft long	$ 85 ppd

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SAY YOU SAW IT IN HAM RADIO

NEW products

get connected — to packet radio

Get Connected to Packet Radio, by Jim Grubbs, K9EI, is the first book devoted exclusively to packet radio operation. Three major sections cover packet radio from the beginning through intermediate levels.

In the first chapters, Jim offers a quick look at packet radio history. For those contemplating the purchase of a terminal node controller for packet radio, information is included on the rapidly growing number of units available. A comparison of several major designs is included.

Once readers assembles the necessary equipment, *Get Connected to Packet Radio* takes them by the hand in a step-by-step process leading to their first successful first packet QSO.

The introductory section continues with information on possible problems, useful commands, and a discussion of high-frequency packet operation versus VHF operation. You'll learn where to find additional packet information in magazines and newsletters and how to contact packet organizations, no matter where you live.

The second section begins with a look at packet protocol. Information on the Xerox 820 computer for use as both a packet terminal and as a bulletin board are included. A special chapter on accessories takes a look at everything from special software to connect alarms.

In other chapters, details on special packet operations (running a bulletin board, for example) are outlined. The final pages include an extensive appendix containing a handy glossary, a bibliography, and lists of pertinent addresses, frequencies, command summaries, and more.

Get Connected to Packet Radio is available from Ham Radio's Bookstore, Greenville, N. H. 03048, for $12.95 plus $3.50 shipping and handling.

MFC sells ham products division

Microwave Filter Company, Inc. has sold its Unadilla/Reyco Amateur Radio Products Division to Ralph H. Jannini of Antennas, Etc., of Andover, Massachusetts. The division produced baluns, traps, switches, and antenna kits that were distributed through approximately 200 dealers nationwide and to exporters in Canada, South America, and Europe. Jannini will add Unadilla/Reyco to his present operations and will continue to serve MFC's Amateur Radio customers.

For information on Unadilla/Reyco products, contact Antennas, Etc., 16 Hansom Road, Andover, Massachusetts 01810.

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PO-1 Pully Kit $8.50
SO-2 $64.50
SO-3 Commercial Antenna Standoff $99.50

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MRF433		12.5W	12.00	30.00
MRF449,/A	Q	30W	12.50	30.00
MRF450,/A	Q	50W	14.00	31.00
MRF453,/A	Q	60W	15.00	35.00
MRF454,/A	Q	80W	15.00	34.00
MRF455,/A	Q	80W	12.00	28.00
MRF458		80W	20.00	46.00
MRF475		12W	3.00	9.00
MRF476		3W	2.75	8.00
MRF477		40W	11.00	25.00
MRF479		15W	10.00	23.00
MRF485*		15W	6.00	15.00
MRF492	Q	90W	16.75	37.50
SRF2072	Q	65W	13.00	30.00
SRF3662	Q	110W	25.00	54.00
SRF3775	Q	75W	14.00	32.00
SRF3795	Q	90W	16.50	37.00
CD2545		50W	23.00	52.00
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2SC2290		60W	15.00	36.00
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MRF221	15W	136-174	10.00	—
MRF222	25W	136-174	14.00	—
MRF224	40W	136-174	13.50	32.00
MRF237	4W	136-174	3.00	—
MRF238	30W	136-174	13.00	30.00
MRF239	30W	136-174	15.00	35.00
MRF240	40W	136-174	18.00	41.00
MRF245	80W	136-174	28.00	65.00
MRF247	75W	136-174	27.00	63.00
MRF260	5W	136-174	7.00	—
MRF261	10W	136-174	9.00	—
MRF262	15W	136-174	9.00	—
MRF264	30W	136-174	13.00	—
MRF607	1.75W	136-174	3.00	—
MRF641	15W	407-512	22.00	49.00
MRF644	25W	407-512	24.00	54.00
MRF646	40W	407-512	26.50	59.00
MRF648	60W	407-512	33.00	69.00
SD1441	150W	136-174	74.50	170.00
SD1477	100W	136-174	32.50	78.00
2N3866*	1W	30-200	1.25	—
2N4427	1W	136-174	1.25	—
2N5591	25W	136-174	13.50	34.00
2N6080	4W	136-174	7.75	—
2N6081	15W	136-174	9.00	—
2N6082	25W	136-174	10.50	—
2N6083	30W	136-174	11.50	24.00
2N6084	40W	136-174	13.00	31.00

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MRF136	21.00	SAV7	30.00
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MRF150	89.50	2SC1946A	12.00
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2N4048	7.20	2SC2289	22.00
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COLLINS STATION 75S3 receiver, 32S1 transmitter, 516-F speaker and P/S, connecting cables, extra clean $600. K6KZT, 2255 Alexander, Los Osos, CA 93402.

WANTED: Meter for Gonset GSB-100 transmitter. Gary Embler, 732 College Avenue, Palo Alto, CA 94306. (415) 424-1483.

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CONTENTS SHACK: List SASE. Ken Hand, WB2EUF, Box 708, East Hampton, NY 11937

DISCOUNT CATV CONVERTERS/DECODERS and Video accessories. Send for free information and prices - it could save you BIG MONEY on your next purchase of these and other CATV items. Easy View, (HR) PO Box 221, Arlington Heights, Illinois 60006. (312) 952-8504. Ask for Rudy Valentine.

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WANTED: Operations manual Hammarlund 105TR. Andy Nagy, 1444 Dill Road, South Euclid, Ohio 44121.

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ATTENTION AMATEURS Send for Free discount catalog. Amateur Communications, 2317 Vance Jackson, San Antonio, TX 78213. (513) 734-7793.

BEAM HEADING CHART, 10 page report in binder with 9 data *fields calculated from your exact QTH to over 540 DX locations.* $9.95 from John Daley, KB6JGH, PO Box 4794, San Jose, CA 95150.

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IBM/APPLE COMPUTER full feature log program, "Hamlog". Send SASE KA1AWH, Box 2015, Peabody, MA 01960.

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THE GOOD SAM HAMS invite RV operators to check in the Good Sam Ham net 14.240 *Sundays 1900Z also 3.880 Tuesdays at 2359Z. Net control N5BDN, Clarksville, IN*

OLD RADIO transcription discs wanted. Any size, speed. W7FIZ, Box 724 HR, Redmond, WA 98073-0724.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $10.00. Beginners RTTY Handbook $8.00. PO Box RY, Cardiff, CA 92007.

IMRA, International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, New York 10538.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

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CHASSIS and cabinet kits. SASE K3IWK, 5120 Harmony Grove Road, Dover, PA 17315.

COMING EVENTS

Activities — "Places to go . . ."

PENNSYLVANIA: The Nittany Amateur Radio Club's annual Mt. Nittany Ham Festival, July 12, Firemen's Carnival Grounds, Pleasant Gap. 8 AM to 4 PM. Admission $3.00 per ham. Spouses and children free. Flea market space $5 includes admission. Program and good food available. Talk in on 146.16/.76 (north and west) 146.25/.85 (south and east) and 146.52. For information: Chuck McMullen, K3CM, 7 Holly Circle, State College, PA 16801.

MARYLAND: BRATS Hamfest and Computerfest, sponsored by the Baltimore Radio Amateur Television Society, Sunday, July 27, Howard County Fairgrounds, Rt. 144 at Rt. 32, West Friendship. Indoor tables with AC power $20 each; 4/$75. Without AC power $10 each; 10/$95 by advanced registration only. Dealer setup Saturday 2 PM. Sunday tailgating space $4 per space. Free VE exams 11 AM Sunday, no advanced registration required. Talk in on 63/03, 16/76 and 52. For information and reservations: W3GXK, BRATS, PO Box 5916, Baltimore, MD 21208.

MICHIGAN: The 38th annual upper Peninsula Hamfest, sponsored by the Delta County Amateur Radio Association, Saturday, August 2, Bay De Noc Community College, Escanaba. Registration $2. Activities, meetings, swap and banquet. For information: Aileen Gagnon, WA8DHB, Co-Chairman, 9159 Bay Shore Drive, Gladstone, MI 49837.

WEST VIRGINIA: The 27th annual West Virginia State ARRL Convention, Friday evening July 4 through Sunday, July 6, Jackson's Mill State 4-H Camp near Weston, Lewis County. Reception Friday evening. Exhibits, tech forums, equipment auction and *pizza party.* Free flea market space for registrants. Plenty of family fun. For information: Sue Goodwin, 103 Cleveland Avenue, Nitro, WV 25143.

KENTUCKY: The Central Kentucky ARRL Hamfest, sponsored by the Bluegrass Amateur Radio Society, Sunday, August 10, 8 AM to 4 PM, Scott County High School, Longlick Road and US Rt. 25, Georgetown. Tickets $4.00 advance; $5.00 at the gate. Tech forums, license exams, awards and exhibits. Air-conditioned facilities. Free outdoor flea market space. For information or tickets: SASE to Scott Hackney, KI4LE, 629 Craig Lane, Georgetown, KY 40324.

PENNSYLVANIA: The 49th annual Hamfest sponsored by the South Hills Brass Pounders and Modulators ARC, Sunday, August 3, south campus of Allegheny Community College, West Mifflin, 7 miles south of Pittsburgh. Talk in on 146.13/73 and 146.52. For information: Doug Wilson, WA3ZNP, 185 Orchard Avenue, Emsworth, PA 15202.

NEW YORK: Electronics Extravaganza sponsored by the Putnam Emergency Amateur Repeater League, Saturday, August 16, J.F. Kennedy Elementary School, Brewster. 9 AM to 4 PM. Admission $3.00. Children under 12 free with adult. Food and refreshments available. Indoor tables, with one admission, $10.00. Tailgating $5.00. FCC exams. Limited walkins. Talk in on 144.535/145.135. For table reservations or information: R. Dillon, N2EFA, RFD 7, Noel Court, Brewster, NY 10509.

PENNSYLVANIA: The Mid-Atlantic Amateur Radio Club's annual Hamfest, Sunday, August 10, 8ucks County Drive-in Theater, Rt. 611, Warrington. 8 AM to 3 PM. Rain or shine. Admission $3.00 plus $2.00 for each tailgate space. Bring your own tables. Talk in on WB3JOE/R, 147.66/.06 or 146.52. For information: MARCBox 352, Villanova, PA 19085 or call Bob Josuweit, WA3PZO (215) 449-9727.

TEXAS: Texas State ARRL and Texas VHF/FM Society's joint convention, July 11-13, El Tropicano Hotel, 110 Lexington, San Antonio. Hospitality suites, licensed day-care, QCWA breakfast, Saturday evening banquet. ARRL forum. For information: Melvin H. Anderson, Jr, 8932 Saddle Trail, San Antonio, TX 78255 (512) 698-1712 or 698-1714.

NEW JERSEY: SCARC '86, sponsored by the Sussex County ARC, July 19, Sussex County Fairgrounds, Plains Road, Augusta. Doors open 8 AM. Registration $3.00. *Indoor tables* $7.00 each. Tailgate space $5.00. Food and refreshments. Free parking. Talk in on 147.90/30 and 146.52. For information: Donald R. Stickle, K2OX, Weldon Road, RD 4, Lake Hopatcong, NJ 07849. (201) 663-0677.

PENNSYLVANIA: The Murgas Amateur Radio Club's 7th annual Hamfest, July 6, Ice-A-Rama in the Coal Street Sports Complex, Wilkes-Barre. Setup 6 AM. General admission 8 AM. Tables available by advance reservation. VE exams. Talk in on 146.61, 53.61 and 146.52. For information or to register for VE exams call (717) 388-6863.

INDIANA: *The WA9SNT Amateur Radio Club will hold its* annual Swapfest, August 9, ITT Technical Institute, 9511 Angola Court, Indianapolis. 8 AM to 4 PM. General admission $2.00. Students $1.00. Flea market space $1.00 additional. Setup time 6 AM. Auction, large flea market, refreshments available. ITT Technical Institute will hold an open house during the Swapfest. All interested are invited to tour the facilities. For additional information: Dave Johnston, K9HDQ, c/o ITT Technical Institute, 9511 Angola Court, Indianapolis, IN 46368. (317) 875-8640.

WISCONSIN: The Eau Claire Amateur Radio Club's annual Hamfest, Saturday, July 12, 4-H Buildings in Eau Claire. 8 AM to 4 PM. Tickets $3.00. Free tables and coffee. *Talk in on 31/91* and 52 simplex. For information/tickets SASE to Gene Lieberg, KA9DWH, 2840 Saturn Avenue, Eau Claire, WI 54703.

WISCONSIN: The South Milwaukee Amateur Radio Club's annual Swapfest, Saturday, July 12, American Legion Post 434, 9327 South Shepard Avenue, Oak Creek. 7 AM to 4 PM. Parking, picnicking, food and refreshments available. Admission $3.00 includes "happy time" with free beverages. The Milwaukee Volunteer Core Group will conduct Amateur Radio exams during the day. For more details, including a map, write: South Milwaukee ARC, PO Box 102, South Milwaukee, WI 53172-0102.

1986 "BLOSSOMLAND BLAST" Sunday, October 5, 1986. Write "BLAST", PO Box 175, St. Joseph, MI 49085.

CALIFORNIA: FCC exams, Novice-Extra. Sunnyvale VEC ARC. (408) 255-9000 24 hour. 73, Gordon, W6NLG, VEC.

PENNSYLVANIA (Meadville) July 5. Firecracker Hamfest by Crawford Amateur Radio Society. Electronics, radios, computers. New location at Meadville Recreation Center just off Rt. 27 E in Meadville. All indoors; no rain out. One olympic and two child-size deluxe swim pools with lockers and showers and other sports available in Center. Free indoor flearmarket space; bring table. Dealers: free indoor space and table(s) available by preregistration; reserve early. Admission adult $2.00, children free, swimming for small extra. Talk in 144.53/145.13, W3MIE/R. For information write CARS Hamfest-86, PO Box 653, Meadville, PA 16335 or call Ben Ferer, KF3F, 814/724-2432.

BRITISH COLUMBIA: Maple Ridge Hamfest, July 12 and 13, St. Patricks Center, 22589 121 Avenue, Maple Ridge. Admission Hams $6.00; non-hams over 12 $3.00. Under 12 free. Two hams in family $9.00. Commercial displays, flea market, food, children's programs. Nearby shopping and recreation areas. Camper space, no hookups. Talk in on 146.20/80 and 146.34/94. For more information: Bob Haughton, VE7BZH, Box 292 Maple Ridge, BC V2X 7G2 or phone (604) 467-4915.

WEST VIRGINIA: The 8th annual TSRAC Wheeling Hamfest/Computer Fair, Sunday, July 20, Wheeling Park. 9 AM to 4 PM. WV's largest. Dealers welcome. 30,000 sq. ft under roof. 5 acres of flea market. Family activities at Park. Admission $3.00 in advance, $4.00 at door. To reserve space contact: Jay Paulovicks, KD8GL, RD 3 Box 238, Wheeling, WV 26003 (304) 232-6796. For tickets, TSRAC, Box 240, RD 1, Adena, OH 43901 (614) 546-3930.

INDIANA: The 16th annual Indiana ARRL Convention and Indianapolis Hamfest, Saturday and Sunday, July 12 and 13, Marion County Fairgrounds, Indianapolis. Featuring the largest electronic flea market and new Amateur Radio equipment display in the state. Gate fees $5.00 with free parking on grounds. Gates open 6 AM. Large covered flea market. Inside tables $8 each. 2m Fox Hunt, homebrew contest. QSL card contest. Free camper hookups available on grounds. Nearby restaurants and motels. For information contact Bill Evans, WB9BEN (317) 745-6389.

MASSACHUSETTS: The MIT UHF Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday, July 23, 7 PM, MIT Room 1-134, 77 Mass Avenue, Cambridge MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-0160/646-1641 or Craigo Rodgers (617) 494-1986. Exam fee $4.25. Bring copy of current license (if any), two forms of picture ID and completed form 610 available from FCC in Boston (223-6609).

OPERATING EVENTS

"Things to do . . ."

OHIO STATE FAIR Special Event, August 1-17. Listen for W8TO 80-10 meters. Exchagge QTH and RST. All correspondence and requests for awards to ARS W8TO, att: State Fair event coordinator, 280 East Broad St, Columbus, Ohio 43215.

The Eastern Michigan ARC will operate K8EPV during the annual Port Huron to Macinac Island Yacht Race, July 19-20. 1400Z to 0200Z both days. For certificate send QSL and SASE to CBA or 654Georgia, Marysville, MI 48040.

High Plains ARC will operate K7YPT at historic Fort Laramie, July 4 and 5. QSL for business SASE to K7YPT, PO Box T, Torrington, WY 82240.

The Texas ARRL and VHF/FM Society convention, Alamo City, W5SC continues the San Antonio Sesquicentennial celebration of Texas independence, July 12 to 13. 10, 15 and 20 meters. QSL. No. 10 envelope, SASE to W5SC, 90 Brees Blvd, San Antonio, TX 78209.

ALBANY, NY, is celebrating its 300th anniversary. The year long activities will peak during Tricentennial Week, July 18-22, 1986. The Albany Amateur Radio Association will be providing public service communication for events and special QSL cards as well as sponsoring the "Worked Albany Radio Members" WARM Award.

OKLAHOMA Amateur Radio operators will conduct their 3rd annual "Field Day" exercises Saturday and Sunday, July 12 and 13, Lake Canton beginning 2 PM Saturday and continuing through noon Sunday in conjunction with the annual IARU Radiosport DX Contest. For information: Tim Mauldin, WA5LTM, Lake Canton Field Day, PO Box 19097, Oklahoma City, OK 73144 (405) 521-5048.

EAST AURORA, NY. July 27. 12th annual Racing Day. Pioneer Radio Operators Society will operate W2OFC to help the community celebrate its heritage as a turn-of-the-century capital of breeding and racing champion horses. QSL with business SASE to W2OFC, 308 Parkdale avenue, East Aurora, NY 14052.

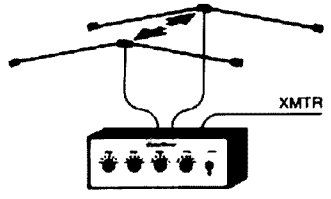
The Illinois Valley ARC will operate special event station KD9UL to commemorate the 25th anniversary of manned spaceflight in conjunction with the Spaceweek event, 1300 UTC to 0100 UTC. For a certificate and QSL, send SASE and your QSL to Bruce Boston, KD9UL, 815 East Third Street, Beardstown, IL 62618.

ACB RADIO AMATEURS a special-interest affiliate of the American Council of the Blind (ACB) will operate special event station KW4U from 0000Z June 29 to 2400Z July 5 from the Hilton Hotel in Knoxville, TN in conjunction with ACB's silver anniversary convention. Submit QSL card confirming QSO with KW4U during convention week and receive an attractive commemorative certificate. Send QSL's to John McCann, KW4U, 2105 N. Illinois Street, Arlington, VA 22205.

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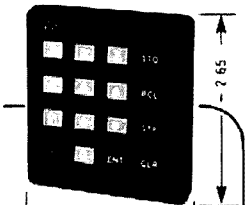
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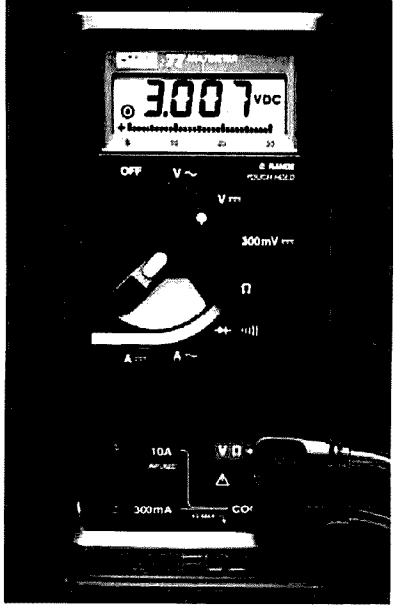
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THE GUERRI REPORT by Ernie Guerri, W6MGI

new technique uses lasers to etch microchips

In the traditional process for manufacturing ICs, as many as 100 individual steps may be required. Each step introduces the possibility of error and increases the risk that the finished product will contain some flaw that renders the chips unusable. Depending on the size and complexity of the chip, less than half of the finished wafer may yield acceptable chips, leaving the surviving chips to recover the entire cost of fabrication.

Not surprisingly, IC manufacturers are constantly seeking ways to improve yield and reduce costs. In one new technique, a laser used as a photoetching device scans the surface of a silicon wafer in the presence of certain gases. Under static conditions, these gases have no effect on silicon; the energy of the laser, however, decomposes the gases into compounds that define active elements and interconnects by either etching away unwanted material or by deposition onto the substrate. The source and drain regions of a transistor are made by doping the silicon with phosphorus, which the laser creates by breaking molecules of phosphine gas. Hydrogen chloride, which serves as an etchant, is activated by the thermal energy of the laser beam. Interconnects on the chip are made by similarly decomposing gases that contain tungsten, nickel, and polysilicon.

One of the major incentives for this new method is a national program, led largely by the Departments of Defense and Energy, to develop new classes of supercomputers. Much of this work has been done at the Lawrence Livermore National Laboratory, where experiments indicate that the technique can produce as many as 1000 transistors per second. At this rate, it would be possible to fabricate supercomputer chips — consisting of about 100,000 transistors each — at the rate of one per day.

Other exciting possibilities include repairing damaged high-value chips and turning a new design into a prototype chip in one day or less, as opposed to today's turnaround time of one to four weeks.

One company (XMR of Santa Clara, California) already offers commercial equipment for this technology. If the technique achieves its promise, we can expect a whole new generation of advanced-capability semiconductors.

"WaferScale" integration

Still on the subject of semiconductors, and one of the major beneficiaries of laser fabrication, is another technique just coming into its own — WaferScale integration. This technique uses the surface of a silicon wafer to implement an entire functional capability. Examples include complete 32-bit microprocessors, with memory and all relevant I/O functions, a "silicon" hard disk with 20Mb of storage, RAM speed, and all disk controller functions on a single wafer.

This technique promises to make very complex functions available in a single package. But this improvement is not without peril. Because of the large amount of circuitry and the extensive processing required on such devices, any mistake in fabrication results in a very expensive piece of scrap. Also, the large number of circuits and functions possible with WSI makes packaging considerations a major concern; it may be necessary to have hundreds of pins on a very complex functional element — more than can now be accomodated. However, the general benefits of WSI seem to justify the complexities of making such devices, and within the next few months the first few WSI products are expected to be announced.

new super-magnet makes smaller motors

An essential component of many motors is the large, heavy permanent magnet associated with the non-excited element of the motor. Magnequench™, a new product developed by the Delco Division of General Motors, is about 25 percent stronger than any other known magnetic material.

Currently, the most widely used high-power magnets, composed of samarium-cobalt, are expensive and difficult to manufacture. But GM's new material is so low in cost that the auto maker plans to use the material in starter motors on some 1986 cars. GM reports that only 5 ounces of the material are needed for the newly designed motors.

This tremendous saving in weight and size offers several benefits. Using a smaller, lighter starter, for example, simplifies design of the engine area. Using Magnequench instead of conventional magnets in all the control motors of a car would presumably produce a measureable effect on fuel economy as well.

Because samarium-cobalt is the material sometimes used to make very small, high-performance loudspeakers, it should be interesting to see if the speakers manufacturers put in our HTs get any better as supermagnets become more widely available.

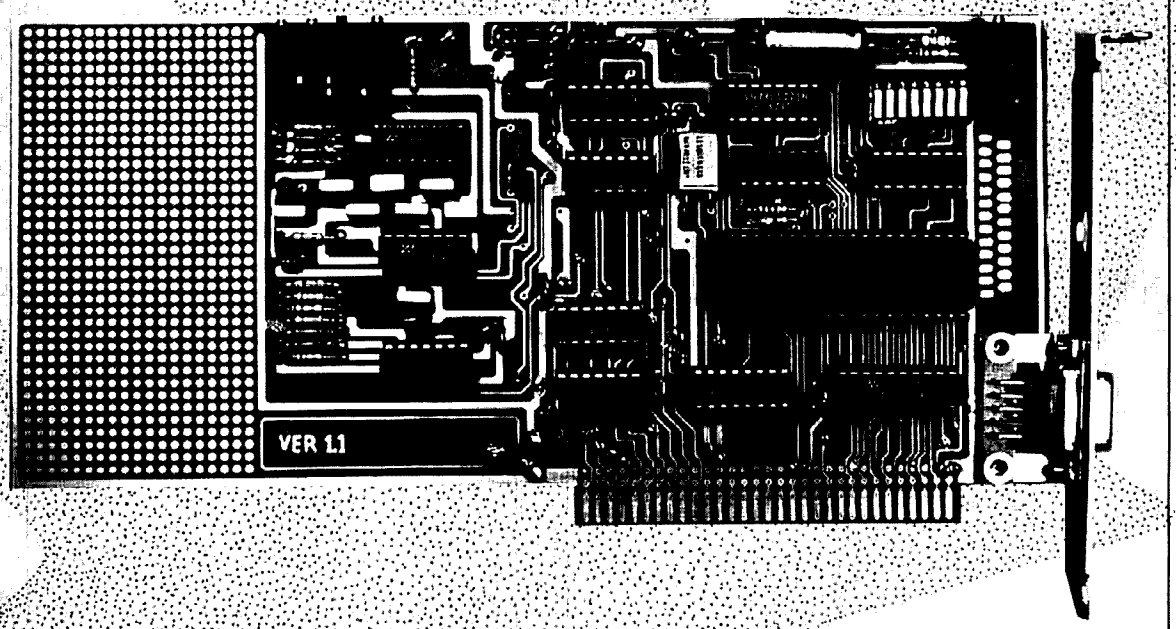
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PACKET RADIO

ham radio magazine

from Canada—



TNC for the IBM PC

focus on communications technology

ham radio magazine

AUGUST 1986

volume 19, number 8

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Rich Rosen, K2RR
editor-in-chief
and associate publisher

Dorothy Rosa, KA1LBO
assistant editor

Joseph J. Schroeder, W9JUV
Alfred Wilson, W6NIF
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Dorothy Sargent, KA1ZK
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circulation manager

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circulation

cover art:
Barbara Smullen

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REFLECTIONS

making history

It **wasn't Chernobyl** — just a few small-town hams meeting over dinner to plan the first use of Packet Radio in an evacuation simulation in the towns surrounding a local nuclear power plant. While packet had been used in similar exercises elsewhere with some success, this was a "first" for our rural area.

Even though the task — handling incoming and outgoing health and welfare traffic — seemed straightforward enough, there were plenty of problems to work out. There was the familiar challenge of finding volunteers for a weekday drill during normal working hours. There were also problems peculiar to packet itself. Given the relative newness of the medium, convincing the planning agencies of the viability of packet hadn't been easy; the drill would be a test not only of the emergency response system as a whole, but of the local Packet Radio system under pressure.

Although there's plenty of packet activity in this part of New Hampshire, it isn't what you'd call a high-traffic area. Digipeaters are few and far between, and hilly to mountainous terrain can make propagation quirky at best. Then there's the complexity of the medium itself: even experienced operators have been known to stare at their CRTs for embarrassingly long moments of uncertainty. (The popular W0RLI software includes an incredulous "What?" response for those occasions when an operator's instructed the packet board to do something incomprehensible or just plain silly. But while the operator may not be willing to admit complete and utter discombobulation, the equipment, or at least the software, is refreshingly frank in this regard.)

The evacuation simulation — and its packet component — were successful. Time-critical traffic was received and passed immediately; less urgent traffic was received, stored on a packet bulletin board, and then forwarded to the National Traffic Service (NTS) or other packet stations well within deadlines.

Packet radio is *here*. It's hot, growing like crabgrass, and becoming both more accessible and affordable by the day. How hot? The first printing of Jim Grubbs, K9EI's book, *Get *** Connected to Packet Radio*, was sold out in only 60 days. Manufacturers say they're just barely keeping up with demand for new equipment. Experienced packeteers upgrading to new equipment have no trouble selling their old TAPR-1's at hamfests. There's no question that younger hams, particularly in urban areas, are turning to packet in impressive numbers. Senior hams maintain a smaller but conspicuous and growing presence.

There is help for newcomers: Jim's book, for example, and instructional videotapes available from Kantronics and TAPR go a long way toward demystifying Packet Radio. And like the legendary Scout who escorted the elderly woman across the street against her will, Packet operators bring missionary zeal to their avocation. It isn't difficult to get them talking about Packet Radio; it *is* difficult to make them stop.

Sometimes we'll hear someone who's read about Packet but not tried it say that it just doesn't make any sense. And sometimes somebody who's tried it will insist that it *still* doesn't make any sense. Packet is new and anything but simple; it has been suggested that if some operators, particularly in high-traffic urban areas, were more patient with newcomers — at least as patient as W0RLI's software, with its endearing "What?" response — Packet Radio might experience less attrition.

It's one thing to get Amateurs into Packet Radio and another to keep them there. Affordable equipment and accessible technology will get them there; patience will keep them.

Dorothy Rosa, KA1LBO
Assistant Editor

presstop de W9JUV

A "NEW" ELECTRONIC COMMUNICATIONS PRIVACY ACT WAS PASSED by the House and sent to the Senate June 12. The new act, HR-4952, is essentially a somewhat tempered rewrite of the bill approved by Rep. Kastenmeier's subcommittee May 14 (see July Presstop). This version, which received House Judiciary Committee approval June 10, is much less "hardline" than the previous HR-3378 which, for example, would have made the mere "interception" of a protected transmission illegal -- even if it were scrambled and thus unintelligible to the person who'd tuned it in! Under this new version, no crime occurs unless there is an "acquisition of the contents" -- a change suggested by the Association of North American Radio Clubs.

The Definition of "Readily Accessible To The General Public" has also been improved and clarified. To be off-limits under HR-4952, a transmission would either have to be scrambled or encrypted, use "secret" modulation techniques, be carried on a subcarrier or via satellite, transmitted by a common carrier (but not paging), or on broadcast link frequencies. Specifically not protected are government, law enforcement, public safety, aeronautical, marine, CD, cordless telephones, and of course, Amateur, CB, and CMRS.

Unfortunately, Even Though This Bill Is Much Less Objectionable than its predecessor, it still suffers from the same basic defects discussed at length in the February, 1986 ham radio editorial. It takes away an established right, that of tuning a radio wherever one wishes, and then restores it piecemeal. And it's still an unenforceable attempt to provide an illusion of privacy (by Congressional mandate) to a private industry, cellular radio.

The Bill Has Now Gone To The Senate Where It's Been Adopted, word-for-word, as S-2575. Though one Senate aide has said he expects the bill "to move pretty quickly" now that it's received House approval, it's sure to receive some critical Senate scrutiny. Nevertheless, full Senate approval could come as early as late July after the Senate returns from its July 4th recess. ANARC has asked the Senate Subcommittee on Patents, Copyrights, and Trademarks to hold at least one additional hearing before bringing the bill to a vote.

Even If This Version Of The Communications Privacy Act Becomes Law in its present form, ANARC, in particular, along with SCAN, several Amateur Radio publishers, and many others who wrote their legislators still deserve most of the credit for making the bill's provisions far less onerous and restrictive than they almost certainly would have otherwise been.

OSCAR 10 IS OFF THE AIR, QUITE POSSIBLY FOR GOOD. Its problem surfaced as a major memory malfunction in the satellite's Integrated Housekeeping Unit (IHU) on May 17, and since then attempts to communicate with OSCAR 10 and perform damage control procedures have had little success. The best hope now is that battery discharge during an early fall eclipse period may permit the IHU problem to correct itself so at least some control can be regained when the solar panels come back into sunlight and recharge the batteries.

Launch of JAMSAT's JAS-1 Satellite Is Still On Schedule for July 31 at 2030 UTC. Real-time launch coverage will be provided by WA3NAN on 3855 and 14280 kHz. JAS-1 has two Mode J transponders, one for CW/SSB and the other providing four channels for Manchester-coded FM digital communications. Uplink for CW/SSB is 145.9-146.0 MHz; downlink will be 435.8-435.9 MHz. Beacon frequency is 435.795 MHz; estimated period 120 minutes.

Two New Russian Satellites, RS-9 And RS-10, are now thought to have successfully passed pre-launch checkout procedures and will probably be launched sometime this summer.

FCC IS MOVING TOWARD IMPLEMENTING LAND MOBILE ON 421-430 MHz near Detroit, Cleveland, and Buffalo. Last fall Land Mobile became the primary 421-430 MHz service within a 50-mile radius of the center of each of these cities, so as secondary users Amateurs below "Line A" (the line marking the area along the Canadian border that protects its land mobile from possible U.S. Amateur interference) must avoid interference with operations in those cities.

ARRL Has Asked The FCC To Require Land Mobile Users in those cities to keep base and satellite receivers within 30 miles of each city's center and use tone-coded squelch, in order to minimize interference from Amateurs operating legally outside the protected area.

THE "ARCHIE AND HIS FRIENDS" AMATEUR RADIO COMIC BOOK is nearly finished, and should be ready for distribution when school opens in September. Developed by Archie Comics, it's aimed at 9- to 15-year-olds and features Archie and his pals sharing the fun of Amateur Radio. Half the funding for the 32-page promotion was provided by the Amateur Radio Industry Group that's been working on boosting Amateur Radio, with ARRL picking up the other half.

RESPONSIBILITY FOR MAINTAINING THE AMATEUR RADIO EXAMINATION QUESTION POOL is likely to pass to the Amateur Radio community as the result of imminent FCC action on that proposal. Though the proposal drew concerns that it could compromise VEC program integrity, or that it would result in a lack of uniformity between different VEC's exams, cutbacks in government spending and the consequent need to reduce FCC staff workload make delegation almost certain.

A "YOUNG HAM OF THE YEAR" AWARD HAS BEEN ESTABLISHED by the Westlink Report to be presented at the "Ham-West" Amateur Radio Convention in Las Vegas each November. Any U.S. licensed Amateur under is eligible, and selection will be based upon his or her contributions to community, nation, communications technology, and/or the promotion of the Amateur Radio service. Yaesu has joined Westlink Report in sponsoring the award, and will provide the winner with a Yaesu transceiver plus transportation to and a room at the convention. Nominations should go to Westlink Report, 28197 Robin Avenue, Saugus, California 91350.



comments

odd antennas

Dear HR:

In his letter to the editor (February, 1986, page 9) C.N. Francis, W0MBP, posed a number of interesting but soluble questions about what he calls "odd antennas." I will attempt to answer his questions in the order he presents them.

His first query deals with what is commonly called the Extended Double Zepp antenna. While the extension principle involved in the Extended Double Zepp is a cost-effective means of achieving gain in the high frequency bands, the antenna does have faults. First, the boresite lobe has poorly suppressed (i.e. less than 10 dB) adjacent lobes at about ±53 degrees.[1] Second, as a non-resonant antenna it will always present a complex impedance at the driving point. At VHF and UHF, where wavelengths represent small physical distances, far more tractable antennas can be built using half wavelength segments. For instance, a two-element, half wavelength, parasitic beam can yield about 5 dB of forward gain, a front to back ratio, and a reasonably clean pattern. Alternatively lengthening the Extended Double Zepp to three half wavelengths and bending it forward slightly about its midpoint will produce a radiator with almost 3 dB of forward gain, a good pattern, and being resonant present a convenient driving point impedance.[2]

Next Mr. Francis describes three antennas which are composed of asymmetric dipoles which he confuses with the Windom Antenna. While this mistake is commonly made it should be noted that there is no relationship between them. The asymmetric dipole[3] has a number of faults, among which are: (1) The pattern is skewed. (2) Due to the lack of symmetry there is a strong tendency to develop antenna currents on the transmisison line. (3) The feedpoint position can't be predicted but must be found experimentally – this can be a difficult job. In addition to the above problems the three arrays described have progressive phase shift across their surfaces. Thus each element will have a different mutual impedance acting on it, and each element will then have unique impedance characteristics. Under these conditions locating the proper driving points is much like adjusting the length of a chair leg.

In his final group of antennas, Mr. Francis uses a hybrid arrangement of a Windom feed and a 225 degree loading section. Probably much of the problem at this point is confusion about Windom principles. Some recitation of history may help.

In the early days of radio, communication was carried on at low frequencies. At these low frequencies (long wavelengths) the propagation mode is surface wave, which mandates vertical polarization. At that state of the art, vertical radiators long enough to achieve first resonance simply did not exist. Thus most antennas were operated as electrically short radiators with all the commensurate problems of poor efficiency and poor current distribution. Somewhere along the line it was discovered that a horizontal conductor, connected symmetrically to the end of the vertical member, would raise the radiation resistance and improve the current distribution on the vertical radiator. This of course was the birth of the "T" antenna. Since the currents in the loading section were equal and opposite, its radiation was suppressed. Later when higher frequencies and ionospheric propagation modes came into use more interest developed in horizontal polarization. In 1929 Everitt and Byrne [4] described a modification to the "T" in which a resonant horizontal section is fed about 13-1/2 degrees off center with a random length of single wire feeder. Radiation is suppressed in the lead about 3 dB, and due to the lack of symmetry useful radiation was obtained from the flattop. This antenna was described in QST[5] by Loren Windom W8GZ, and inherited his name. Unfortunately Mr. Francis misses the point that the superstructure must be of resonant length and not 225 degrees. What are described in the letter are radiating single wire transmission lines, top loaded with pathological superstructures. For instance in the BBBC arrangement radiation from the top half of the loading section would be cancelled by the out of phase radiation from the lower half.

In short, all of these antennas could be made to radiate in some fashion, however, they all lack practicality, which could be solved by using simpler and more conventional structures.

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Robert B. Sandell, W9RXC
Urbana, Illinois 61801

short circuit

VHF/UHF world

References 8 and 9 were omitted from W1JR's July column. They should read as follows:

8. Joe Reisert, W1JR, "VHF/UHF World; Stacking Antennas - Part 1", *ham radio,* April, 1985, page 129.
9. Joe Reisert, W1JR, and Gary Field, WA1GRC, *RF-CAD Version 3.6 × : Computer-Aided Design Package for Radio Frequency and Microwave Circuits and Antennas,* available from Ham Radio's Bookstore, $39.95 plus $3.50 postage and handling.

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a packet radio TNC
for the IBM PC

Single plug-in board extends usefulness of popular pc to digital communications

By Jack Botner, VE3LNY, Ron Bradshaw, VE3IUV, Max Pizzolato, VE3DNM, and John Vanden Berg, VE3DVV, Hamilton and Area Packet Network, Box 4466, Station D, Hamilton, Ontario, Canada L8V4S7.

The advent of the IBM Personal Computer, with its open and well-documented architecture, has made it possible for hobbyists to extend the use of the microcomputer into many varied and challenging areas. This article describes a complete packet radio terminal node controller which can be implemented on a single circuit board that plugs into one of the slots in the PC. All that's required to get on the air is to connect up an Amateur Radio transceiver (see **fig. 1**).

the TNC: basic packet building block

Traditionally, packet radio has been achieved through the use of a device called a *terminal node controller*, or TNC. A TNC consists of a processor (such as the 8085 or Z80), a serial or parallel port to connect the TNC to a terminal or microcomputer running a terminal emulation program, and a synchronous port for the communications channel. A certain amount of RAM (random access memory) is available, and the necessary programming is provided on ROM (read only memory).

Because a TNC contains many components that are already present in our microcomputers, merging the function of the TNC into the microcomputer is an effective way to reduce the cost of a packet radio system. With suitable programming, the function of the TNC can be implemented in the microcomputer at a fraction of the cost of a separate TNC. There's a bonus, too, in that a higher level of function is available to the user because of the close coupling of the channel to the host processor.

In this implementation, an Intel 8273 programmable SDLC/HDLC protocol controller chip is interfaced to the bus in the IBM PC. This can be built on a prototype card or on a custom-printed circuit board. It's also possible to construct, on the same card, a simple but effective 1200-bps Bell 202-compatible half-duplex modem, a device in widespread use among packet radio enthusiasts. Software that runs the standard AX.25 packet radio protocol is available. The adapter is called the HAPN-1 adapter; HAPN stands for "Hamilton and Area Packet Network," the name of our packet radio club.

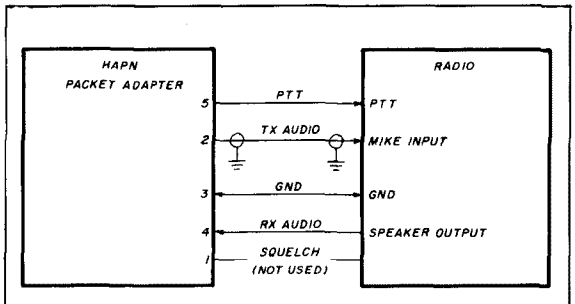


fig. 1. Simple radio hookup via microphone plug and speaker output. Note: Jumper J2 should be installed and J1 removed on adapter board.

hardware

The circuit shown in **fig. 2** uses an 8273 (U6) interfaced to the PC bus and a 1200-bps modem which can be connected to an Amateur Radio transceiver. The

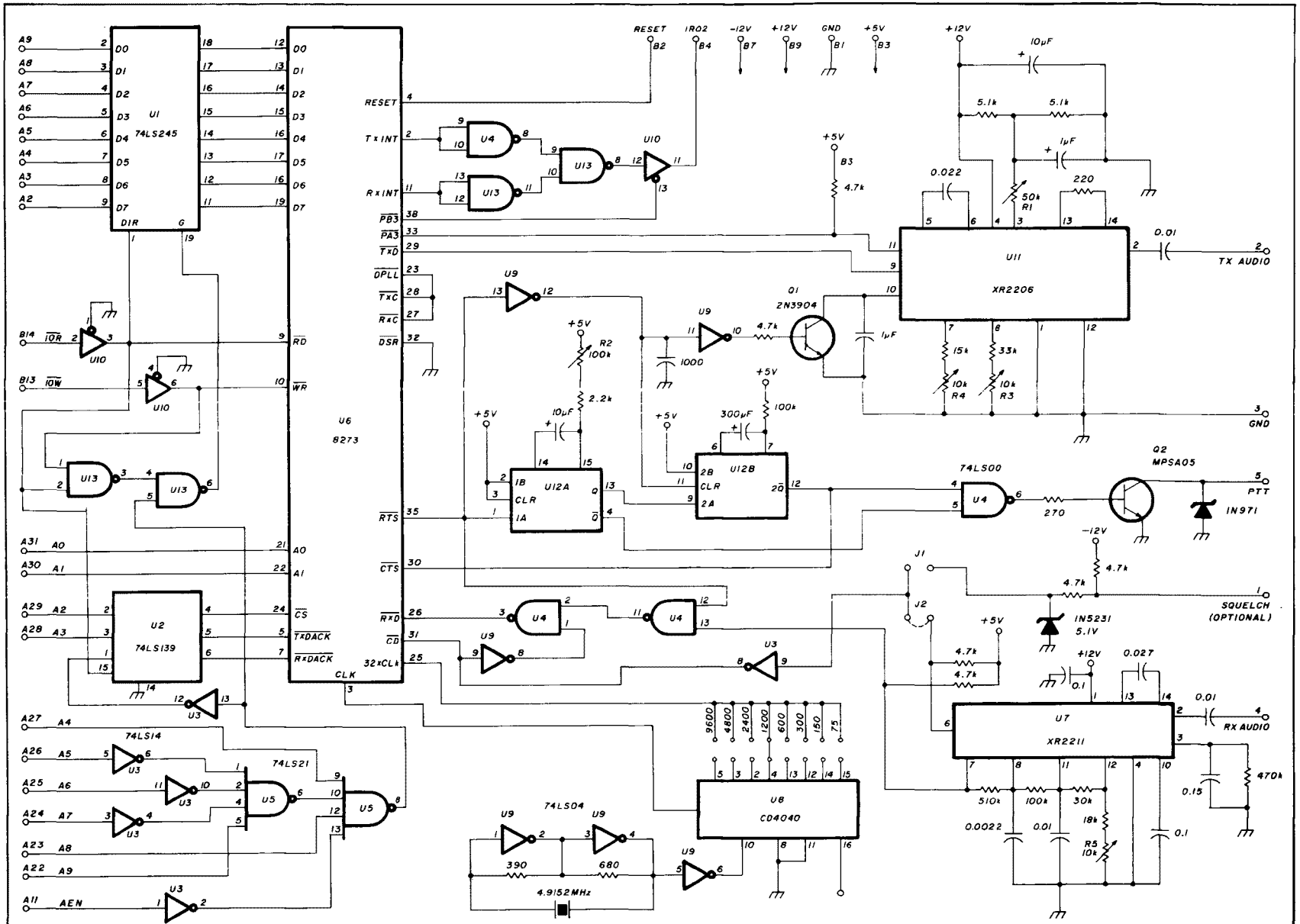


fig. 2. TNC adapter schematic.

8273 chip, called the *protocol controller*, is the most important part of the adapter.

The 8273 data lines are connected to the PC bus using a 74LS245 three-state buffer U1. The state and direction of the 74LS245 are controlled by the I/O Read and I/O Write lines and the I/O decode logic. The I/O decode circuit, made up of U2, part of U3, and U5, responds to addresses in the range of 310 through 31F (hexadecimal notation). The 8273's address requirements are met by a 74LS139 two- to four-line decoder (U2).

Data transfers to and from the 8273 may be done using polling, interrupts, or Direct Memory Access (DMA). This design uses interrupts, because they're easy to implement and permit a form of background operation that doesn't require running the packet application program constantly. The 8273 provides separate interrupt signals for transmit and receive (Tx Int and Rx Int, pins 2 and 11, respectively). The two interrupt lines are ORed together (part of U13) so that the adapter uses only one of the PC's hardware interrupt lines. The software can easily distinguish between transmit and receive interrupts by reading the 8273 status register. The combined interrupt line is buffered to the PC bus with a 74LS125 three-state bus buffer (part of U10) and is wired to the PC's IRQ2 interrupt line. IRQ2 was chosen because it is the least frequently used interrupt line in the PC. The circuit allows the interrupt line to float on the bus until the 8273 is initialized and the PB3 control port on the 8273 is activated. This means that the card can share IRQ2 with other hardware as long as the other device also floats its IRQ2 line when not in use and the two devices are not used at the same time.

clock signal

The 8273 features a digital phase-locked loop which makes it possible to derive the synchronous clock signal from the receiver data stream, a feature that greatly simplifies the design of the hardware. Pins 27 and 28 of the 8273 (Tx Clock and Rx Clock) are tied together to pin 23 (DPLL output). The 32X clock signal is provided by a 4040 binary counter (U8), giving a selection of data rates from 75 to 9600 bps by jumper or dip switch. A 4.9152-MHz crystal oscillator (U9) provides the clock signal for the 8273 and U8, the baud rate divider.

modem

The modem uses a pair of Exar chips, the 2206 function generator (U11) and the 2211 FSK demodulator (U7). The 2206 generates the two tones, 1200 and 2200 Hz, used for transmit; the 2211 decodes the two tones from the receiver audio. Two control signals are required by the 8273: Carrier Detect (CD, also known as DCD or RLSD in various contexts), indicating that the radio channel is busy, and Request to Send (RTS), indicating that the 8273 wants to transmit.

channel busy indication

The CD signal (channel busy indication) can be derived two ways, either internally by using the data carrier detect (DCD, pin 5) from the 2211 or externally by using the squelch from the receiver. Using the squelch to indicate the radio channel busy condition is superior but usually requires modifications to the radio. It guarantees, however, that the system will not transmit when another signal — possibly voice or some other non-data signal — is present on the channel. The internal DCD from the 2211 signals only when actual data is being received. This is the recommended way when no squelch line is available or the QRM level is high, such as 20-meter operation.

To use the internal DCD, install jumper J2. The 2211 checks the incoming signals for data and signals DCD to the 8273. To use the squelch CD, install jumper J1. The two resistors and zener diode convert the squelch signal from the radio to a TTL level. The signal is debounced by using a 74LS14 Schmidt trigger inverter gate (part of U3). An example of the squelch pickup from the radio is given in **fig. 3.**

transmit mode

To transmit, the 8273 brings up RTS, which triggers U12a (CTS-delay single shot) and brings up PTT via U4 pin 5 (negative or gate) and the 2MPS-A05 switching transistor (Q2). The U12a single shot allows the radio to stabilize when going into transmit before sending data; this function is called the *clear-to-send delay*. When U12a times out, the watchdog timer U12b starts, bringing up the CTS line for the 8273. This signals the 8273 to send the data. At the same time, it continues to activate the PTT line via U4 pin 4. When the 8273 is finished transmitting it drops RTS, which applies a reset to the watchdog single shot, causing it to drop the PTT line. If for any reason the RTS line stays up too long (as in a crash of the system, for example) the watchdog times out after about 22 seconds and shuts down the transmitter, releasing the radio channel to other users. Transistor Q1 enables the tones on the 2206 only when RTS is up (i.e., when 8273 is transmitting), so that there's no interference audio from the modem when the microphone is used for normal voice conversation.

Figure 1 shows how the adapter can be hooked up to a handheld radio. The TX-audio line is connected to the microphone audio input by means of a shielded cable. The receiver audio and PTT are taken from the speaker plug. Remove jumper J1 and install it at J2 to use the DCD line from the demodulator. For best performance, adjust R1 (a 50 k trimpot) for about 3-kHz deviation.

Integrated circuit voltage and ground pinouts.

IC	Type	Gnd	+5	+12
U1	74LS245	10	20	–
U2	74LS139	8	16	–
U3	74LS14	7	14	–
U4	74LS00	7	14	–
U6	74LS21	7	14	–
U6	8273	20	40	–
U7	XR2211	4	–	1
U8	CD4040	8	16	–
U9	74LS04	7	14	–
U10	74LS125	7	14	–
U11	XR2206	12	–	4
U12	74LS221	8	16	–
U13	74LS00	7	14	–

TNC to 2-meter radio interconnections

Figure 3 shows a hookup to an IC22s 2-meter radio. Here the TX-audio is inserted after the audio pre-amp, Q29, in the IC22s. If a DTMF (touchtone) input is available, it could be used instead. A 390-ohm resistor takes off the RX-audio before the volume control. The PTT goes to the front panel microphone connector. Transistor Q100 inverts and buffers the radio squelch. All the connections are brought out via the utility connector on the back of the radio. An audio output of about 1000 mV will give you about 3-kHz deviation on transmit. Remove jumper J2 install it at J1 on the packet adapter board. Set the CTS delay (R2) for about 300 msec; if it's set too short, the beginning of the transmitted frame will be missing and therefore impossible to copy. If it's too long, time will be wasted transmitting idle flag characters.

combining TNC with the host computer simplifies system design

In the traditional TNC, the programming is supplied on a ROM IC chip, and is relatively invisible to most users. In the HAPN implementation, the software is supplied as programs on a diskette, which are run like any other program on the PC.

One of the limitations of the traditional TNC is the loose coupling, or distance, between the TNC and the microcomputer acting as the terminal. By necessity the TNC strips the packet data down to the most elemental level, that of a serial stream of characters. This makes it difficult for the host computer to learn anything about what's going on in the TNC at the packet level. On top of this is placed a layer of commands, escape characters, and flow control, making even the best of packet host programs appear clumsy and unfriendly.

But including the packet hardware in the PC itself overcomes all of these problems. The programmer is free to design the software — particularly the end-user interface — in any way desired, without the need to compromise for the TNC interface. The result is a high-function packet system with a friendly user interface.

HAPN packet software implementation

In order to make effective use of the 8273, the following must be included:

- a BIOS (Basic Input/Output System)-like hardware driver that can be loaded as an extension to the system to drive the 8273;
- a protocol manager with an application program interface; and
- an application program that is run at the discretion of the user.

In addition, the 8273 must be made to operate in the background so that the node represented by the 8273 can be active all the time.

These requirements can be met by developing the software in at least two programs. The first program, M25, contains the hardware driver, protocol manager, and application program interface. The second program, C25, makes up an end-user application program. Since the 8273 operates on interrupts, no application program assist is necessary to receive or transmit data. Total independence from the application program has been achieved by using the PC's timer tick hardware interrupt exit as the protocol manager's dispatcher for scheduling events.

M25, the hardware driver module, contains a number of distinct functions and is made up of several routines. These routines have been link-edited to make up one program, which is run once after the system has been booted to become a resident operating system extension.

The application program, C25, contains a set of end-user functions primarily concerned with the sending and receiving of data. It can be used to activate the hardware functions in M25 via the application program interface. C25 would be run any time the user wishes to access the packet node; otherwise the PC can be used to run other application programs.

The main function of M25 is to service interrupts from the 8273, so that data can be transferred into and out of the channel. The interrupt handling routine is divided into two parts, one that handles transmit and one that handles receive interrupts. During its operation the 8273 generates interrupts for each byte transmitted or received, as well as for various transmit and receive frame complete events. Also a variety of error conditions may be reported to the processor — for example, when a frame has been received with an invalid CRC (cyclic redundancy check) field, generally caused by noise.[1]

The receive interrupt handler stores each byte of a frame in a buffer. When the frame is complete the 8273 generates another interrupt. At that point the frame may be inspected to make sure it's free of errors and is valid by the rules of protocol in use. Frames that

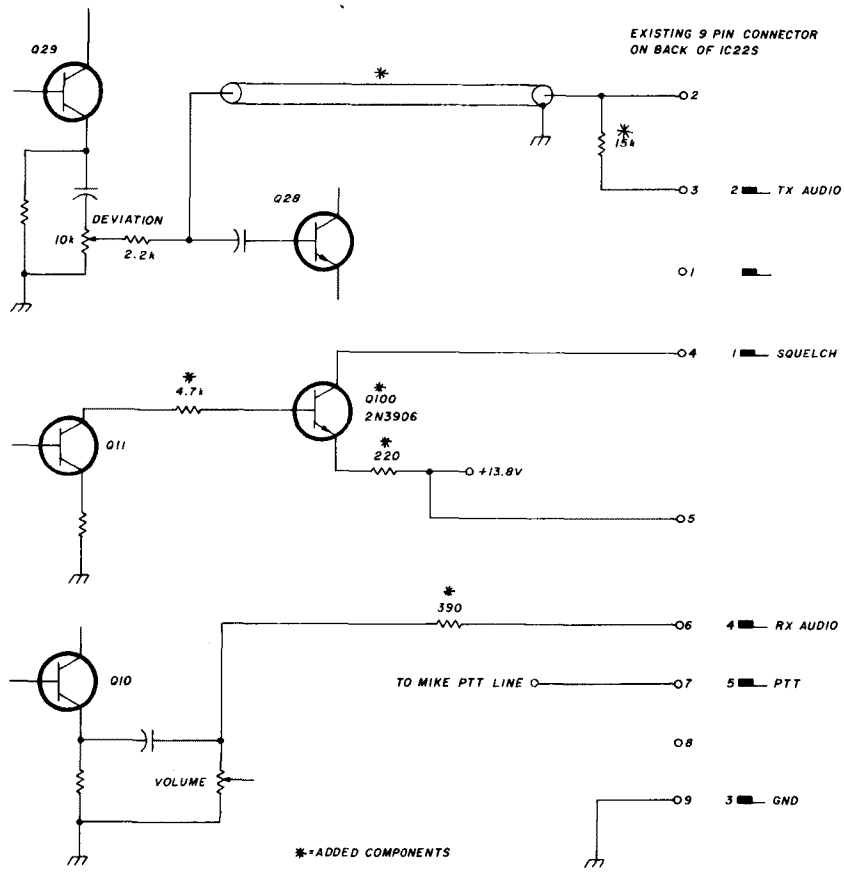


fig. 3. Example of squelch pickup.

contain errors are simply discarded. If a valid frame is received, some link-level protocol management is performed on the frame before exiting the interrupt routine. An example of this processing would be to look for connect requests and queue a response when one is received. It's also possible to delete request frames from the transmit queue after a valid response has been received.

The transmit interrupt routine passes each byte of the frame from a buffer to the 8273 as required. When the frame has been completely transmitted, the 8273 generates an early frame complete interrupt. At this time the software can decide if another frame should be transmitted in the same packet. In the AX.25 protocol, up to seven frames may be transmitted in one packet. If another frame is available, a command is issued to the 8273 to transmit another frame during the current transmission. The 8273 automatically inserts a byte of flags between frames sent in this way.

When the packet is complete the 8273 generates a transmit complete interrupt. The software uses this interrupt to do some protocol management on the transmit queue. For example, because request frames may have to be retransmitted at a later time, they're left on the transmit queue until they're acknowledged; response frames, on the other hand, are transmitted only once and may be deleted from the transmit queue at this time.

In this implementation an OR gate is used to drive both transmit and receive interrupts on the same IRQ line to the PC's 8259 interrupt controller. M25 distinguishes between them by reading the 8273's status register, which contains two bits that specify which type of interrupt was generated. It's interesting to note that because of noise on the channel, the 8273 occasionally becomes confused and generates interrupts that cannot be identified by the status register. When this kind of error occurs, M25 issues a software reset command and then re-initializes the 8273 to ensure correct operation. Statistics are kept on the various events that occur in the packet system, and may be displayed by the user.

TNC adapter parts list (fig. 2).

R1	50k trimpot
R2	100k trimpot
R3	10k trimpot
R4	10k trimpot
R5	10k trimpot
R6	5.1k
R7	5.1k
R8	220
R9	15k
R10	33k
R12	2.2k
R13	100k
R14	1k
R11, 15, 16, 18, 19	4.7k
R17	470k
R20	510k
R21	100k
R22	30k
R23	18k
R24	390 ohms
R25	680 ohms

All resistors 5 percent 1/4 watt
10-turn trimpots recommended for R3, R4, and R5

CR1	1N5231 (5.1 volts 5 percent zener)
Xtal	4.9152 MHz
Q1	2N3904
Q2	MP5-A05
J1/J2	2 pin jumper connector
J3	8-position DIP switch
DB-9	(male) printed circuit board connector
U1	74LS245
U2	74LS139
U3	74LS14
U4	74LS00
U5	74LS21
U6	8273
U7	XR 2211
U8	CD 4040
U9	74LS04
U10	74LS125
U11	XR 2206
U12	74LS221
U13	74LS00

I.C. Sockets: Seven 14-pin
Four 16-pin
One 20-pin
One 40-pin

C1	10 μF tantalum
C2	330 μF tantalum
C3	.001 μF
C4	10 μF tantalum
C5	.022 μF
C6	1 μF tantalum
C7	.01 μF
C8	1 μF tantalum
C9	.1 μF
C10	.15 μF
C11	.027 μF
C12	.0022 μF
C13	.01 μF
C14	.1 μF
C15	.01 μF

M25 contains a dispatcher routine that runs periodically. Its purpose is to manage timeouts and initiate events as they're required by the protocol. The timer tick interrupt, which is issued every 55 milliseconds on the PC, is a convenient way to give control to the dispatcher for this purpose. The dispatcher monitors the channel for activity and initiates frame retransmissions when timeouts occur. A random number generator produces the timeout value so that repeated collisions with another station can be avoided. The dispatcher also keeps track of the number of retransmissions and produces a loss of contact condition when a preset limit is reached. Because it's driven from the timer hardware interrupts, the dispatcher runs independently of any programs on the PC.

M25 also contains the application program interface. This interface allows a program running on the PC to access the 8273 channel. Interface functions include open, read, write, inquire, modify, and close. Note that the application program issuing the open does not have to close before terminating; this means that the packet node will continue to operate regardless of what program is run next on the PC. An alarm feature is also included so that the user is alerted if a connect request is received. The application program interface uses a software interrupt, level 84 (hex), for access by programs.

M25 includes an installation routine for loading itself into memory and becoming resident. Storage is reserved for transmit and receive buffer pools and a trace table. The application program interface vector is loaded so that the 8273 driver interface can be accessed. However, packet operation doesn't begin until it's initiated by an open request from an application program such as C25.

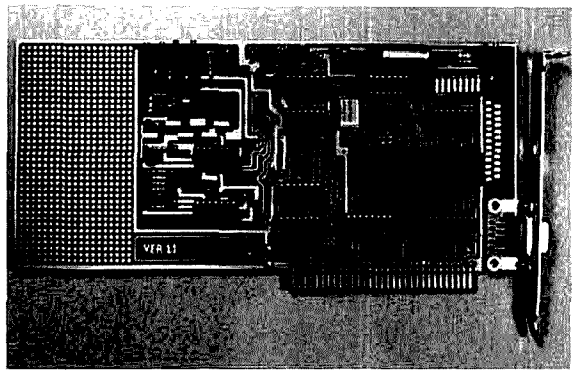


fig. 4. HAPN adapter board plugs into IBM PC slot.

C25, the end-user application program, is an ordinary DOS program designed as an interface between the user (keyboard, display) and the packet channel. It tests to see if the 8273 driver is installed and issues various requests to the 8273 driver application program interface.

data entry

The user of a packet node is mainly interested in sending and receiving data. This data may be entered through the keyboard or read from a disk file. Received data is displayed on the screen and may be saved to a disk file if desired. Other control functions include initiation of a connect request and specification of repeater linking information.

viewing the screen

The screen is divided into three areas: one for displaying received data, another for displaying typed

input, and a third for indicating status. The receive data display area, the largest of the three, scrolls when it fills up. The typed-input area is two lines (160 characters) long. No data is transmitted until the return key is pressed, and a comprehensive set of editing functions is provided so that the user gets an opportunity to change the typed data before it's transmitted. The status area shows information such as lock key status, active function keys, and node callsign and linking repeater information.

other user-friendly features

C25 includes several convenient features. One, the scroll lock key, temporarily stops the screen from scrolling. Another automatically generates, upon request, a test message containing your callsign, the date, and time.

C25 also contains a self-customizing routine so that the program options can be set by each user. Program options are stored inside the program module rather than in a separate file, making C25 more convenient to run. A pop-down menu is provided for entering and displaying repeater information.

There are two exit options from C25: one issues a close to terminate packet operation and one does not. In the latter case the packet node is left operational while other programs are run on the PC. When C25 is run again later it will pick up the status of the packet node using an inquiry call to the application program interface.

Two utility programs are provided in addition to the basic functions described above. S25 formats either the packet trace table or the packet statistics counters. The trace table is useful for analyzing events such as protocol violations, in which two nodes appear to misbehave for no apparent reason. T25 is a test program with a number of functions for testing and setting up the 8273 adapter. It can be used to test the 8273 and bus interface hardware and to perform adjustments on the modem.

building the adapter

The adapter shown in fig. 2 can be constructed on a readily available prototype board for the IBM PC. Consult the technical reference for details of the board's edge connector.[4] Although any suitable connector may be used, a DB-9 male connector was used for the transceiver interconnection. The only critical components are the frequency control trimpots in the modem, R3, R4, and R5. It's a good idea to use multiturn trimpots for these components. You should also use stable mylar or polyester capacitors in the circuits around the 2206 and 2211.

When connecting the adapter to a transceiver, be careful to use good quality shielded wire for the transmit audio lead. It's also a good idea to bypass each end of the cable with a 0.001 μF capacitor and place a ferrite bead over the center conductor at each end. This will prevent RF pickup, which can cause erratic operation of the system and transmission of garbled signals, into the transmit audio circuit of the transceiver.

It's good practice to bypass the +5-volt supply lead to each IC with a 0.1 μF tantalum capacitor. Otherwise, layout and wiring are not critical, if good construction practice is used. For those who prefer not to wire their own boards, both bare boards and assembled and tested boards are available from HAPN.

adjusting the adapter

After the board has been assembled and checked for shorts, it's ready for testing. The first test is an internal loopback test of the 8273 using T25.EXE, the test program supplied with the software. This tests the data buffers (U1), the 8273, address decoding, and the clock circuit.

Now adjust the modem. Start by setting up the two transmit frequencies, F1 and F2. Float U11 pin 10 to enable the tones. Then float U11 pin 9 and adjust R3 for 1200 Hz (use a frequency counter or an accurate scope). Next ground U11 pin 9 and adjust R4 for 2200 Hz. If the output is too low for the frequency counter, adjust the level control, (R^1).

Next adjust the CTS delay by putting a scope on U12 pin 13 and key up the transmitter (using C25.EXE). Adjust R1 for 100-500 milliseconds, depending on how fast your rig switches from receive to transmit; a value of 300-400 milliseconds works well for most rigs.

Demodulator adjustment is a bit more tricky. You have to jumper the connecter pin 2 (Tx audio) to pin 4 (Rx audio) and float U11 pin 10 to enable the tones. When C25 has initialized the 8273 it will transmit flags as 1's and 0's. Adjust R5 so the waveform at U7 pin 7 looks the same as on U11 pin 9.

The remaining adjustment is of the audio level (R1) going to the radio. The level varies widely with the type of hookup into the radio. Adjust this level so the deviation is approximately 3 kHz on the transmitter. This completes the adjustments.

a note on HAPN

The following items may be ordered from HAPN. All prices are in United States dollars and include postage and handling.

- Software only (four programs and documentation), $40.
- Software and bare board (includes set-up program and assembly instructions), $75.
- Software and assembled and tested board (see fig. 4), $199.

Please add $3 for overseas orders.

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Orders should be addressed to HAPN, Box 4466, Station D, Hamilton, Ontario, Canada, L8V4S7.

Dedicated to furthering the state of packet radio, HAPN is currently involved in the design of a station node controller. The station node controller is capable of unifying a packet radio network by providing message routing, gateways, and high-speed links to other station nodes.

HAPN is also involved in experimentation with 4800-baud modems and with V3, the newly developed VADCG protocol. HAPN has experimental V1 and V2 software that runs on the adapter described in this article; for more information, contact HAPN at the above address.

references

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2. J. Botner, "A Packet Radio Adapter for the IBM PC," *QEX: The ARRL Experimenter's Interchange*, January, 1985.
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annunciator bell
for the Kantronics KPC-2

No more missed connections on packet radio

Some computers or terminals can't respond to the bell command. So if you've used one of these machines in a packet radio setup, you may find yourself missing connections with other stations — unless, of course, you're willing to stay glued to the screen watching for attempts to connect.

When a successful connect is made, the Kantronics KPC-2 responds by sending three CNTRL-G ($07 hex) ASCII characters to the terminal or computer. The CNTRL-G character combination is commonly used as the bell command to produce an audible signal. But obviously this works only if your terminal or computer is equipped with the hardware needed to produce an alarm. My computer, a Xerox 820, has no internal bell circuitry. As a result, I was missing connections. By the time I noticed that someone had tried to talk to me, they were often long gone. After only a few weeks on the air I discovered that I wasn't alone in this experience; other packet operators with similar hardware were missing connects, too.

solving the problem

It took only three hours and less than $10 to find a simple solution to the problem. The circuit shown in **fig. 1** doesn't respond directly to the bell character; instead, a piezo buzzer fires for about half a second as the KPC-2 enters a connected condition. There are some advantages to this; my buzzer, for example, will not respond to hidden bell commands that may be embedded in text or files received by my station and the annoying end-of-margin alarm produced by some computers and terminals is also neatly avoided.

circuit

The KPC-2 has a connect status indicator on the front panel. This LED is lit only when a connected link status exists. Data line P-22 from the 63B03 processor drives a 74HC04 inverter to power this indicator. The added circuitry shown in **fig. 2** and external to the KPC-2, uses this output to activate the piezo buzzer to audibly announce the connect status. The alarm is limited to about a half second's duration by a CMOS timer circuit. The exact time may be varied by changing the values of either the 68k resistor or 5 μF capacitor. Note that two inverter stages are used to drive the buzzer. Although they may at first appear to be superfluous, they're needed to produce a well-defined off/on transition for the bell.

The transducer operates from 3 to 20 VDC. Despite being driven directly from a CMOS output, it may provide more volume than necessary. Adding a resistor, up to about 47k, in series with the transducer will reduce the volume. A manual off-on switch removes power from the bell for silent late-night operation.

construction

Table 1 lists the required materials. Assembly is done using point-to-point wiring techniques on the IC perfboard. Be sure to use a socket for the No. CD4001. Don't insert the IC until after all wiring is completed. Note that this board won't fit into the specified enclosure unless some of the phenolic material is carefully trimmed from one of the 2.83-inch edges. I used a nibbling tool for this task, and then dragged the board over the surface of a large file to smooth off

By Peter J. Bertini, K1ZJH, 20 Patsun Road, Somers, Connecticut 06071

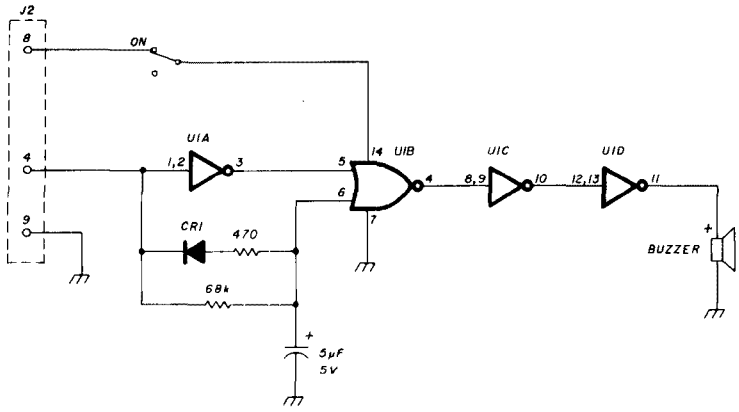


fig. 1. Circuit for the Kantronics KPC-2 bell annunciator. Parts values not specified in the schematic are shown in table 1. The duration of the bell timer may be varied by changing the values of the 5 μF capacitor or 68k resistor. Diode CR1 is a silicon device; a 1N4148, 1N914, or 1N4001 may be used.

Table 1. Bell annunciator parts list.

part	*description*	*Radio Shack part number*
U1	*quad 2-input NOR gate CD4001*	*276-2401*
buzzer	*piezo buzzer, 3-20 volts DC*	*273-065*
case	*4 x 2 x 0.844 inches*	*270-220*
perfboard	*2.83 x 1.85 inches*	*276-150*

the rough edge. After the board is assembled and inserted into the enclosure, be sure to drill a hole directly above the buzzer to make its output more audible.

bell-to-KPC-2 interface

This is the tricky part: chances are your KPC-2 is less than a year old. *Performing this modification will void your warranty.* Note also that this circuit will work only with the KPC-2 hardware. Versions of the KPC-1 — regardless of the software revision level — use an MC6803 processor which does not support the connect-status indicator.

I used J2, the nine-pin D-connector located on the rear apron of the KPC-2, to bring out 5 VDC and connect signals to operate the bell. Pins 4 and 8 were unused. A jumper wire was connected from pin 6 of U14 to pin 4 of the D-connector. Another short jumper ties the regulated 5-volt output from VR-1 to pin 8 of the connector.

Because of possible production changes, manufacturing assembly parameters change, be sure to verify — before work begins — that these pins are unused in your particular KPC-2! I found a jumper, not shown on the schematic, from the pin 4 circuit trace to ground that I had to remove.

A two-conductor shielded cable should be used to interconnect the bell device to the TNC. The plastic hood (supplied with the J2 connector and harness assembly from Kantronics) will have to be removed, reassembled and carefully drilled out to enlarge the cable passageway to allow the bell cable to enter.

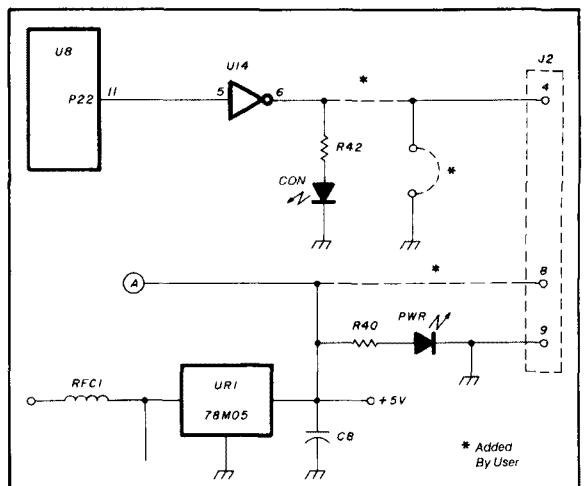


fig. 2. Partial schematic of the KPC-2 shows the wiring added to pins 4 and 8 of connector J2. The connect signal is fed to pin 4, Ground is already present on pin 9, and the 5-volt VCC bus is tied to pin 8. Note the presence of a jumper that originally grounded pin 4; this jumper must be removed from the PC board.

operation

Carefully check your wiring for errors before applying power to the TNC. If everything looks good, hook it all up and connect either to a local ham or to yourself via a local digipeater. A short, pleasant "beep" should reward you as the connect is made.

ham radio

low-noise phase-locked UHF VCO

part 2: construction and testing

This VCO's noise level is over 60 dB below the carrier — great for microwave applications

In part 1 of this article I described what I believe is an answer to the noise problem in voltage-controlled oscillators and showed that most Amateur UHF VCO applications can probably be satisfied by using a single-frequency, dedicated UHF phase-locked loop. This part presents construction information, describes testing, and provides performance characteristics.

construction and circuit characteristics

The UHF PLL is built on a standard 4-1/2 x 6-1/2-inch G-10 (glass epoxy) PC board, as shown in **fig. 1**. The edge connector shown in **table 1** provides for 15 volts DC, frequency select, and an unlock circuit. The latter drives an LED out-of-lock warning indicator. The RF output connector, an inexpensive SMA type,* is soldered to the PC board. Another SMA connector can be added if an external reference signal is to be used.

Oscillator and phase detector. The UHF oscillator and its buffer-amplifier stage are mounted on the trace side of the board with the component side acting as the ground plane. These circuits are enclosed in a simple shield box soldered to the PCB. The box is made of 0.5-inch wide X 0.015-inch thick brass strip, which is available at most hobby shops. The strip is formed into a 1.75 X 2.75-inch rectangle. Before soldering the box to the board, cut slots with a nibbling tool wherever PC lines enter or leave the box area. A partition is soldered inside the box to shield the oscillator from the buffer. An aluminum cover is fitted over the shield box.

The phase detector uses a pair of Schottky diodes. Several kinds of diodes that will work in the phase detector are listed in **table 2**. The traces on the board will accept any of these types, but the Alpha D5486 is recommended as the first choice because it is a dual diode in one package and therefore costs less than a pair of diodes.

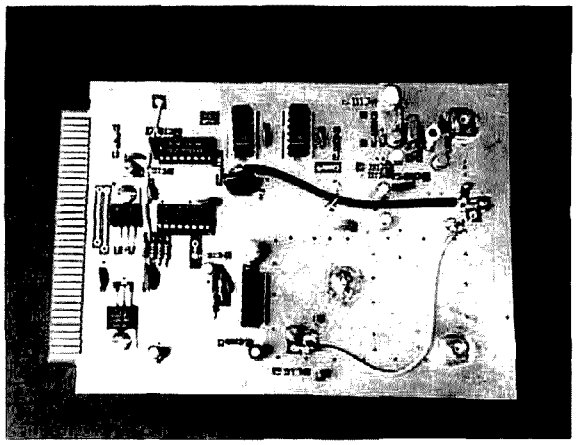
The PC board includes five integrated circuits: an RCA CA3179 prescaler, a 74LS191 programmable divide-by-N, a CD4046 phase/frequency detector, and a pair of 74LS93 reference dividers. Also included are two voltage regulators, a 7808 and 7805.

The quartz crystal for the reference oscillator was obtained from ICM under the code number 471360. This is a shear mode, third-overtone AT-cut crystal. The code number includes information relating to ambient operating temperature, frequency range, type of holder, and operating load. The crystal oscillator coil is a Coilcraft vertically-mounted 12-turn helical "Unicoil" with a 6-32 threaded tuning slug, 1/4 inch long, for inductance adjustment. The inductance without the slug is approximately 0.28 μH. You can wind your own on a surplus coil form if it includes a tuning slug. For example, a 3/16 inch OD coil form 1/2 inch long with a 6-32 powdered iron slug (1/4) inch requires ten close-spaced turns of No. 26 enameled copper magnet wire.

Plated-through board. The PC board is quite special because it is solder-plated and plated through. The through-plating feature is a must for the UHF circuits. Not many Amateurs or experimenters have facilities for through-plating, but there is an alternate approach

By Norman Foot, WA9HUV, 293 East Madison Avenue, Elmhurst, Illinois 60126

* Type 142-0298-001, available from E.F. Johnson, Waseca, Minnesota 56093.



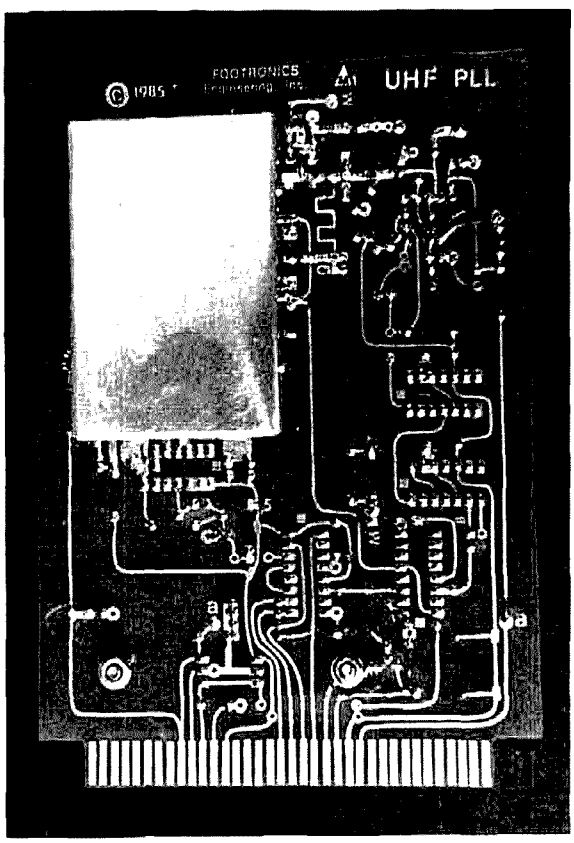


fig. 1[A]. UHF PLL pc board component/groundplane side. [B]. UHF PL pc board trace side. Enclosure in upper left hand corner contains VCO, buffer and loop amplifier circuits.

that I have used with good success. I insert a 0.094-inch long x 0.043-inch diameter brass eyelet in each ground hole. Then I roll the eyelets over with a fine punch, hammer them flat, and solder them to ground on each side of the board. While this is admittedly a rather tedious procedure when there are

table 1. Edge connector terminals.

1 gnd	19 D
2 nc	20 A
3 nc	21 B
4 nc	22 gnd
5 nc	23 unlock
6 nc	24 +5
7 nc	25 +15
8 nc	26 nc
9 +15	27 nc
10 gnd	28 nc
11 +8	29 nc
12 gnd	30 nc
13 +5	31 nc
14 nc	32 nc
15 nc	33 nc
16 nc	34 nc
17 nc	35 nc
18 C	36 gnd

table 2. Diodes that can be used in the phase detector.

Alpha	D5486	1 required
Alpha	D5845	2 required
Alpha	D4787	matched pair
HP	5082-2818	matched pair
HP	5082-2816	matched pair
MaCom	Ma41513	2 required
MaCom	Ma4882	matched pair

over 100 ground holes, it works very well. . . if you have the patience. I use this approach for preliminary breadboards.

The cost of having a single plated-through PC board fabricated at a commercial plant is very high — as much as $230.00. However, in quantities of 100, the price drops to less than 20 percent of this figure. If you're interested in obtaining a board for your project, get in touch (just be sure to include an SASE). If there are enough inquiries, I may be able to provide a board at a very reasonable price.

Avoiding pitfalls. Before heating up your soldering iron in anticipation of assembling this PLL, there are some subtle characteristics that you'll want to know about.

While there's nothing unusual about the component side of the PC board, the trace side includes a number of chip resistors as well as some chip capacitors, particularly inside the oscillator and buffer compartments. To be more exact, the trace side of the board contains 11 chip resistors and 12 chip capacitors.

The chip resistors are used where threading a pair of pigtails through the board from the ground plane

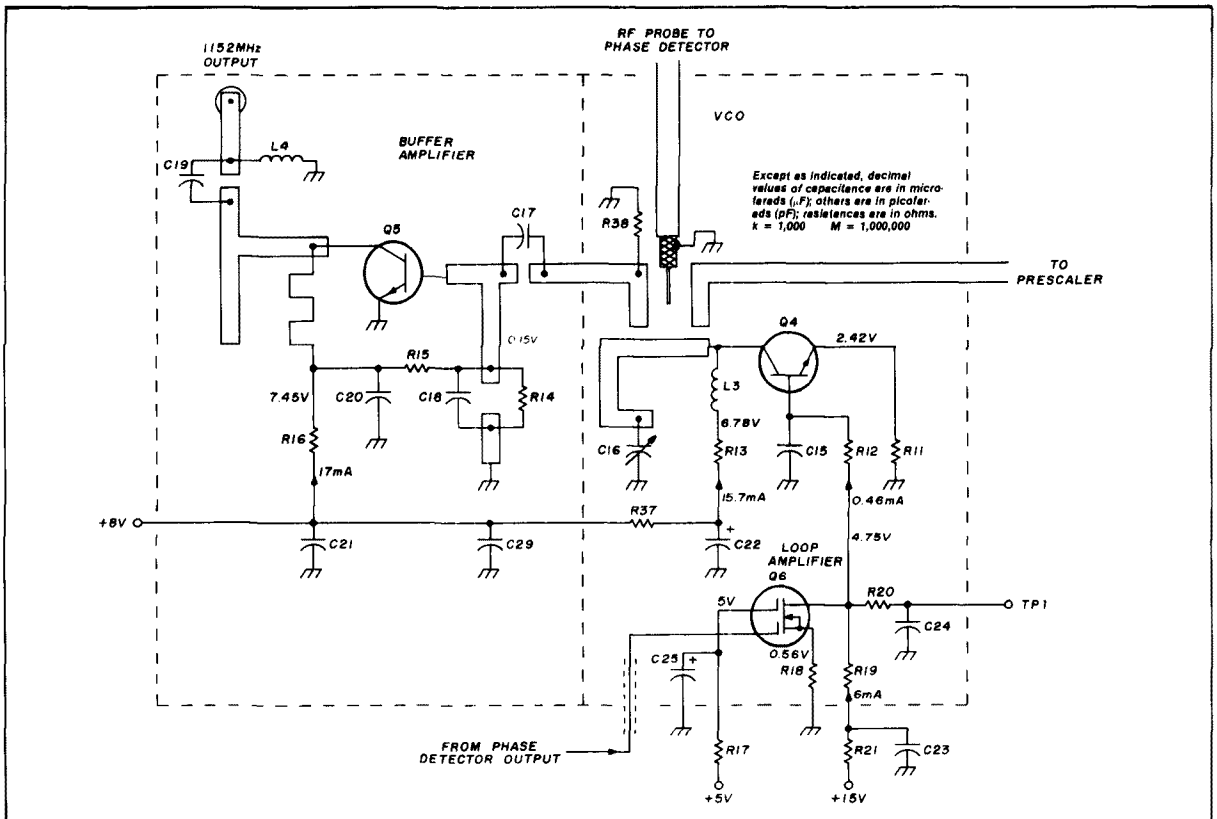


fig. 2. VCO, buffer and loop amplifier schematic.

side would seriously degrade performance. This is particularly true if the resistor is part of an RF circuit, such as an RF termination. Parasitic inductances are close to zero for these devices, which makes their use mandatory in UHF applications.

Chip resistors are now available at reasonable price — typically 6-1/2 cents each in lots of 100 — in standard values from 2.2 ohms to 2.0 megohms. Chip resistors are generally specified at 5 percent tolerance, but since they're laser-trimmed, they're usually more accurate than that. My chip resistors, purchased from Mouser Electronics in Santee, California, are 60 x 130 mils and 0.025 inches thick, or slightly larger than a typical chip capacitor, and easily installed.

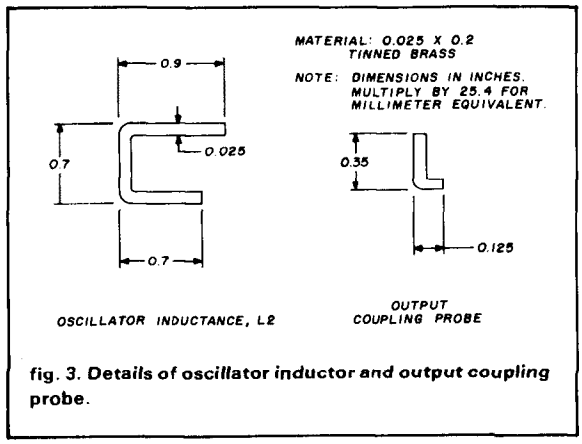


fig. 3. Details of oscillator inductor and output coupling probe.

the oscillator circuit

An article by G. D. Clock and associates[1] describes a means for tuning a VCO by varying the base-emitter current. Television Receive Only (TVRO) and other oscillators have used this principle for a number of years.

Base-emitter tuning. This technique involves adjusting the oscillator base-emitter current to produce frequency changes. Base-emitter and collector capacitances increase with base current. I have produced high-to-low frequency tuning ratios of over 12-1/2 percent by using this method at frequencies as high as 5 GHz. The 1152-MHz VCO described here uses base-emitter tuning, which avoids the need for some-

times difficult-to-find UHF varactor tuning diodes. The oscillator circuit is shown in **fig. 2.**

The base-emitter driving circuit for the oscillator is also shown in fig. 2. The dual-gate MOSFET is a relatively quiet, inexpensive device. It represents the loop amplifier shown in **fig. 4** (see part 1, page 37). TP1 is included to monitor the loop amplifier output voltage for tuneup.

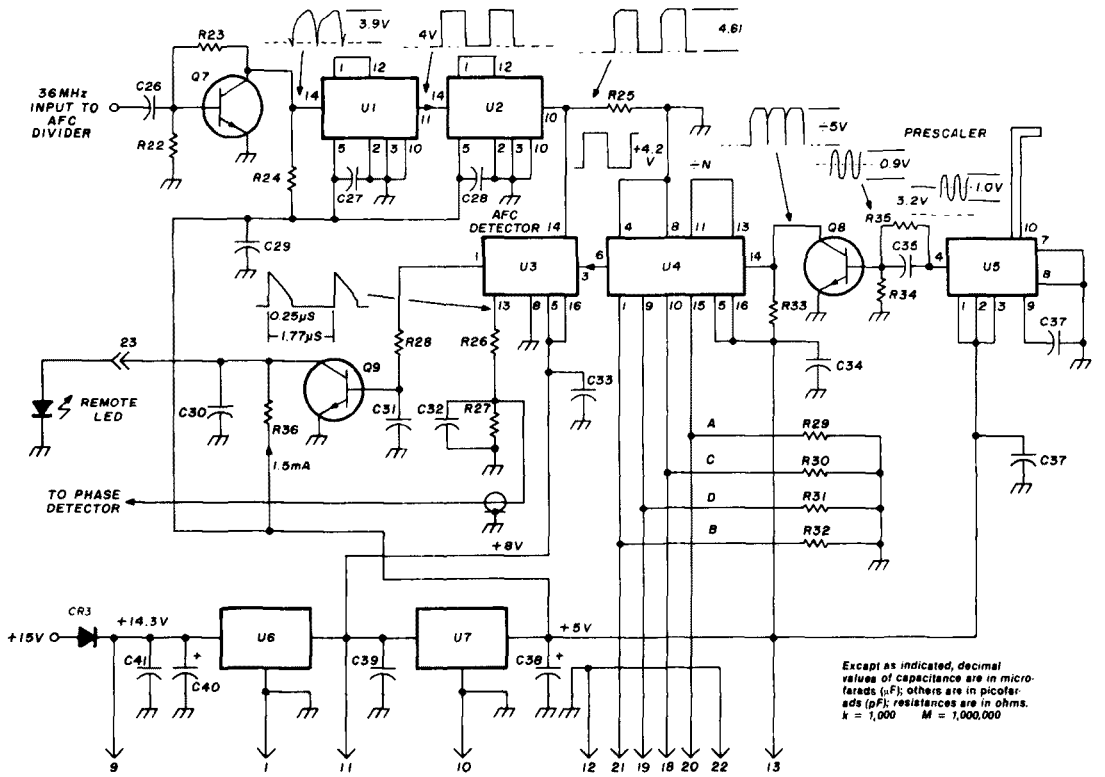


fig. 4. Automatic Frequency Control schematic.

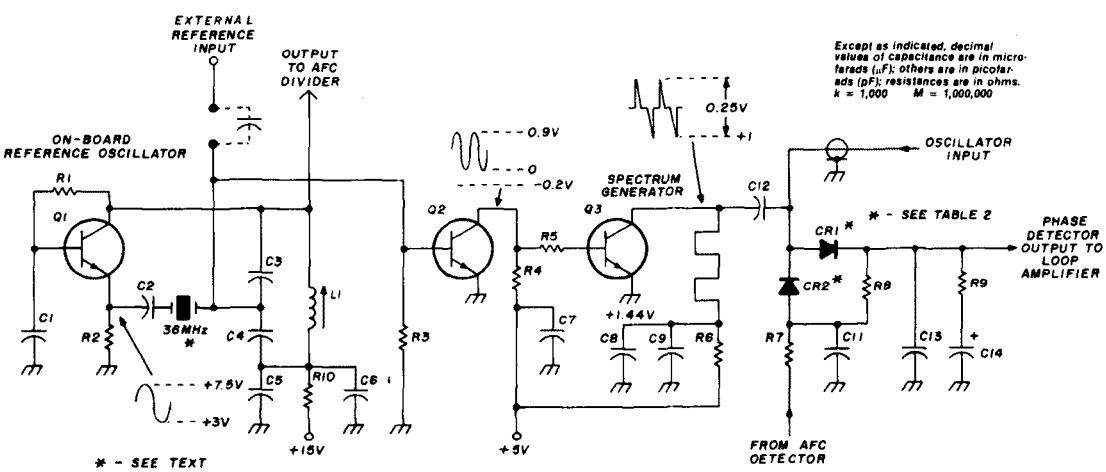


fig. 5. Reference oscillator, spectrum generator and phase detector schematic.

Parts list

C1	51 pF disc
C2	4.7 pF disc
C3	51 pF disc
C4	100 pF disc
C5	0.001 disc
C6	22 μF/35-volt
C7	15 μF/35-volt Tantalum
C8	10 μF/16-volt
C9	1000 pF chip
C10	10 μF/16 volts
C11	1000 pF chip
C12	22 pF chip
C13	47 pF chip
C14	1.0 μF/35-volt Tantalum
C15	4.7 pF chip
C16	Erie No. 560-013, 0.5-5 pF chip
C17	47 pF chip
C19	470 pF chip
C19	47 pF chip
C20	470 pF chip
C21	1000 pF chip
C22	1.0 μF/35-volt Tantalum (inside oscillator compartment)
C23	2.2 μF Tantalum
C24	1000 pF chip
C25	1 μF/50 volts
C26	5 pF disc
C27	0.1 μF 10-volt disc
C28	0.1 μF 10-volt disc
C29	0.1μF 10-volt disc
C30	0.0l μF disc
C31	0.001 μF disc
C32	0.1 μF mylar
C33	0.1 μF 10-volt disc
C34	0.1 μF 10-volt disc
C35	0.001 μF disc
C36	47 pF chip
C37	1.0μF/35-volt Tantalum
C38	100μF/16 volts
C39	0.1 μF 12-volt disc
C40	22 μF/25 volts
C41	0.1 μF 50-volt disc
CR1	See table 2
CR2	See table 2
CR3	1N4002
L1	crystal oscillator coil
L2	oscillator inductance
L3	deleted
L4	5 turns No. 26 tinned busbar, 1/16-inch diameter, 1/4-inch long
PCB	printed circuit board shield box and cover
Q1	2N2369A
Q2	2N2369A
Q3	AT-41435-5
Q4	MRF-901
Q5	AT-41435-5
Q6	40673
Q7	2N2369A
Q8	2N2369A
Q9	2N2222
R1	33 kilohms
R2	820 ohms
R3	22 kilohms
R4	6.8 kilohms
R5	51 ohms chip
R6	680 ohms chip
R7	150 kilohms, 1/8 watt or chip
R8	47 kilohms, 1/8 watt
R9	2.2 kilohms chip
R10	100 ohms
R11	150 ohms chip
R12	4.7 kilohms chip
R13	22 ohms 1/8 watt
R14	910 ohms 1/8 watt
R15	6 kilohms, 1/8 watt (adjust value for IC20 mA if necessary)
R16	33 ohms
R17	2.2 kilohms
R18	100 ohms chip
R19	1.5 kilohms, 1/8 watt (inside amplifier compartment) chip
R20	4.7 kilohms chip
R21	100 ohms
R22	15 kilohms
R23	8.2 kilohms
R24	330 ohms
R25	1 kilohms
R26	10 kilohms
R27	10 kilohms
R28	4.7 kilohms
R29	1.2 kilohms
R30	1.2 kilohms
R31	1.2 kilohms
R33	330 ohms
R34	150 kilohms
R35	39 kilohms
R36	330 ohms
R37	51 ohms (in oscillator compartment) chip
R38	51 ohms (in oscillator compartment) chip
U1	74LS93
U2	74LS93
U3	CD4046
U4	74LS191
U5	CA3179
U6	7808
U7	7805

Frequency adjustment. The VCO uses an MRF-901 transistor. Except for minor details, the circuit is similar to the one used in a previous article.[2] A buffer amplifier provides isolation between the oscillator and the outside world, thus making the job of the PLL easier. It also provides more output power. The buffer employs an Avantek AT-41435-5 bipolar transistor.** An Erie 560-013 piston trimmer, used to adjust frequency, is located at the far end of the line where a tuning varactor might otherwise be located. Although ceramic trimmers also work well, some of them tend to seize after extensive use; the Erie trimmer is more expensive, but can take more abuse. This is an adjustment that needs to be made only once. Thereafter, small changes in oscillator circuit parameters will be compensated for by the AFC circuit.

Coil and coupling probe. To allow the VCO to be used for other purposes, the oscillator "coil" is suspended above the PC board ground plane rather than being printed. This also provides higher *Q*, which is needed to maintain low phase noise. The inductance for 1152 MHz is formed out of 0.025-inch thick by 0.2-inch wide brass strip bent into a U-shape as shown in **fig. 3**. It's mounted by soldering one end to the piston trimmer and the other to the MRF901 collector lead. The collector lead is bent up at right angles to meet the coil.

Coupling to the buffer is accomplished by means of an L-shaped probe located near the transistor end of the oscillator inductance. A 50-ohm chip resistor terminates the buffer input circuit to avoid instability. The coupling probe is an L-shaped bracket made from the same kind of material as the inductance and soldered to the end of the buffer input PC trace. Space the probe about 1/8 inch away from the oscillator induc-

** Available from Peak Technology, Inc., Arlington Park, Illinois

tance. An identical probe drives the prescaler, with very light coupling.

The oscillator's emitter is left unbypassed. This imposes some negative feedback, which in turn tends to suppress low-frequency flicker noise in a manner discussed in reference 9 (see part 1, page 38).

Frequency division. Referring to **fig. 4**, note that a

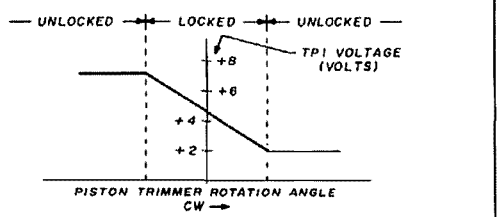


Fig. 6. AFC tuning curve.

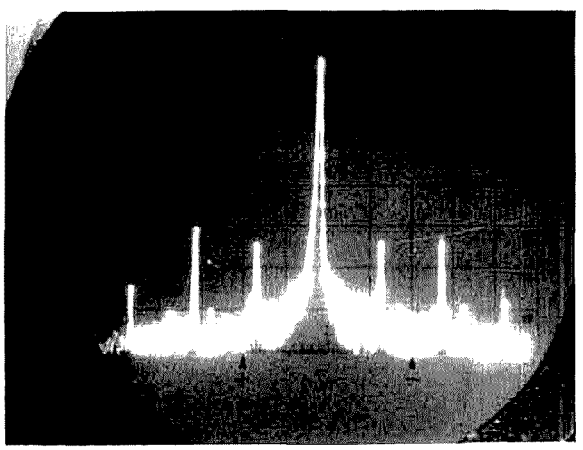


fig. 7. Oscillogram of VCO spectrum with the 4046 as phase detector. Horizontal = 30 kHz/cm; vertical = 10 dB/cm.

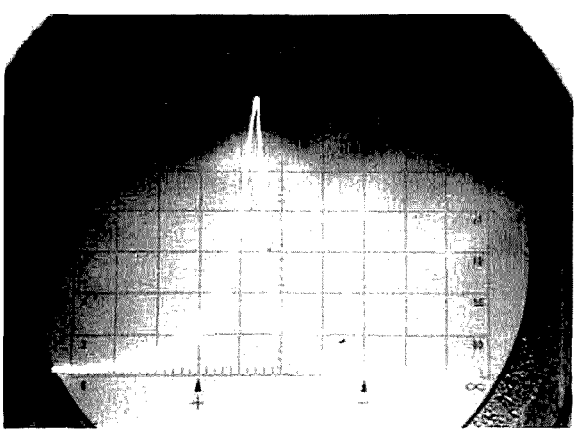


fig. 8. PLL spectrum. Vertical = 10dB/cm; horizontal = 30kHz/cm; BW = 1kHz.

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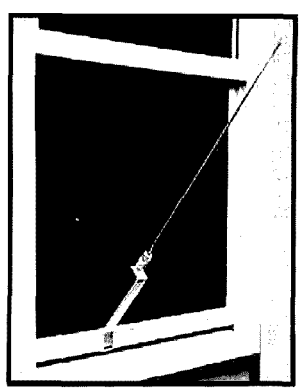
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pair of 74LS93s connected in cascade divide the crystal frequency by 64, to 562.5 kHz. The prescaler, U3, together with the divide-by-N stage, U4, provides division by 2048, which is also 562.5 kHz when the VCO is phase-locked to 1152 MHz. These two signals represent the scaled-down reference and VCO that drive AFC detector U3.

The circuit of U4 is somewhat unusual because the RC output resets the counter to a preprogrammed value. This creates a modulo-N circuit that will divide by any integer from 1 to 16; the most useful numbers are 7, 8, and 9. For 1152 MHz, N = 8.

The AFC output from U3 is fed over a shielded wire to the DC return of the phase detector as shown in **fig. 5**. This wire is located on the component side of the board and its shield is soldered to the ground plane.

Phase detector inputs. Q3 is a UHF bipolar transistor used as the spectrum generator. Like the RF buffer, it's an Avantek type 41435-5. A "picket-fence" of harmonically-related signals, spaced 36 MHz apart across the band from 36 to 1400 MHz or more, drives the phase detector. One of these harmonics represents the desired phase-detector signal which, for an 1152-MHz VCO, is the crystal oscillator's 32nd harmonic.

The phase detector also requires a sample of RF from the oscillator. While initially the VCO frequency may be incorrect, the AFC circuit will steer it to 1152 MHz to effect phase lock. Coupling from the oscillator is achieved over a short (1-1/2 inch) piece of miniature coaxial cable such as RG-174 or equivalent. The braid is stripped back on each end and soldered to the ground plane at the points of entry. One center conductor end is soldered to the input of the phase detector; the other end acts like a probe, extending about 1/2 inch into the oscillator compartment. The insulation is left on the probe end.

noise and spurious sidebands

The original crystal oscillator multiplier built in 1969 included a cavity resonator at the step-recovery diode (SRD) output. The *Q* of this cavity is relatively low

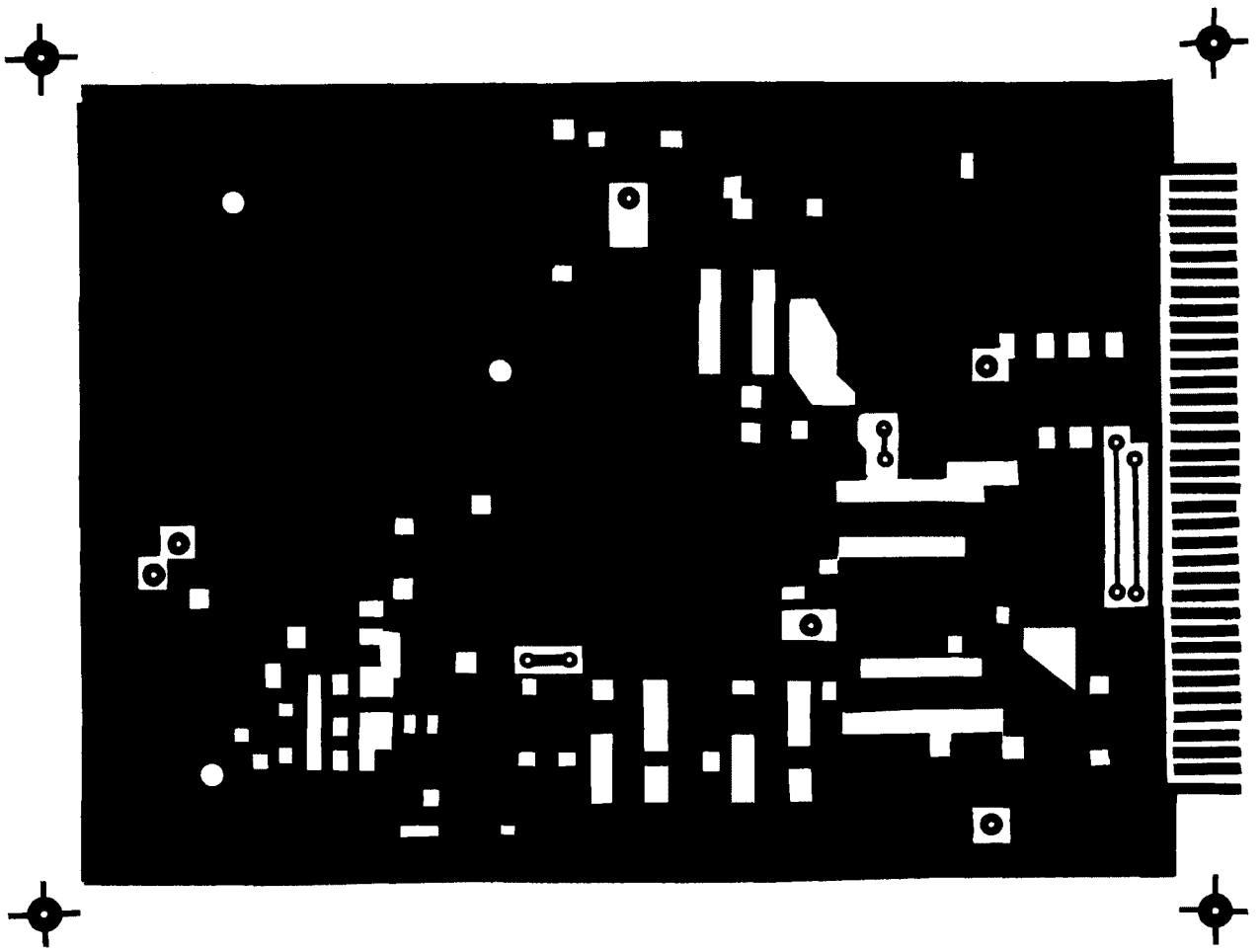


fig. 9(A). Completed UHF PLL double-sided printed circuit board artwork Groundplane side.

because it was tightly coupled in an effort to get maximum power out. Filtering of unwanted sidebands is therefore limited. The strongest set of sidebands appears 384 MHz away from the carrier on each side (384 MHz is the SRD driving frequency). These sidebands are 28 dB below the carrier; all others are over 30 dB below. This represents reasonable performance for narrowband purposes when the first IF is less than 384 MHz, but the spurious sidebands may produce "birdies" at the output of the second mixer in a wide-open (block down-conversion) system.

The UHF PLL described in this article has filtering qualities resulting from the loop filter instead of a cavity. The only significant coherent spurious signals are located 562.5 kHz from the carrier on each side and are over 60 dB below the carrier. No phase noise was detectable on the HP-8551 spectrum analyzer. The calculated level is −145 dBc/Hz at 1 kHz and beyond.

preliminary adjustments

Before applying power, check the component side of the board for insulated holes that don't have components in them. Ignore unfilled plated-through ground holes. All the other holes, with the exception of four pairs of component-side jumpers, should be filled. Then proceed as follows:

- Connect a 100-ohm, 2-watt resistor in series with the 15-volt power supply to limit the current in case of a short; otherwise the current should be about 90 mA.
- Remove the 100-ohm resistor and monitor the current to the board. Depending on the status of tune-up, the current should be between 190 and 230 mA.
- Using a CRO, monitor the waveform on pin 14 of UI. (It's best to use a 50-MHz oscilloscope for this measurement.) Tune the crystal oscillator coil for maximum peak-to-peak voltage. The positive peak should

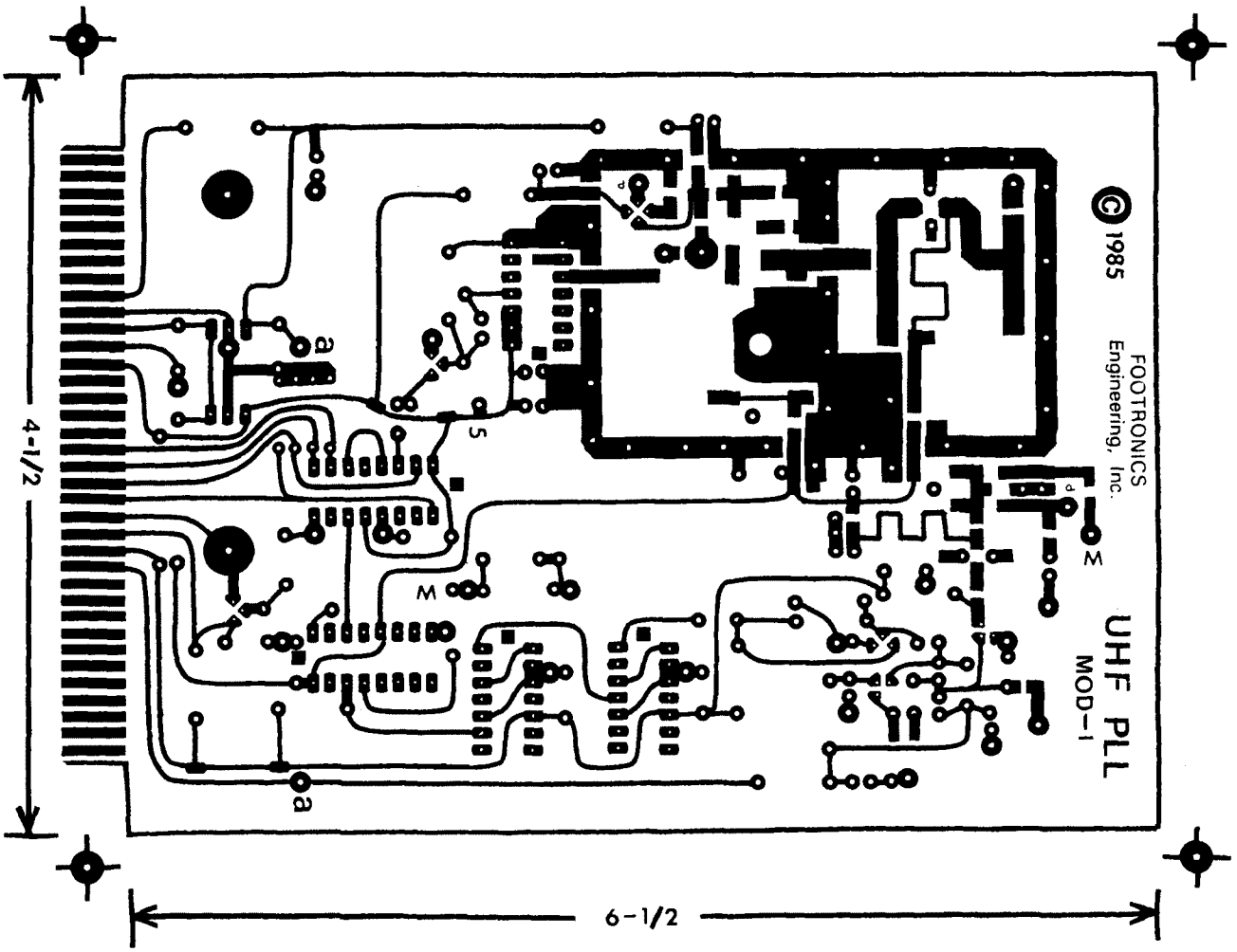


fig. 9(B). Completed UHF PLL double-sided printed board artwork Trace side.

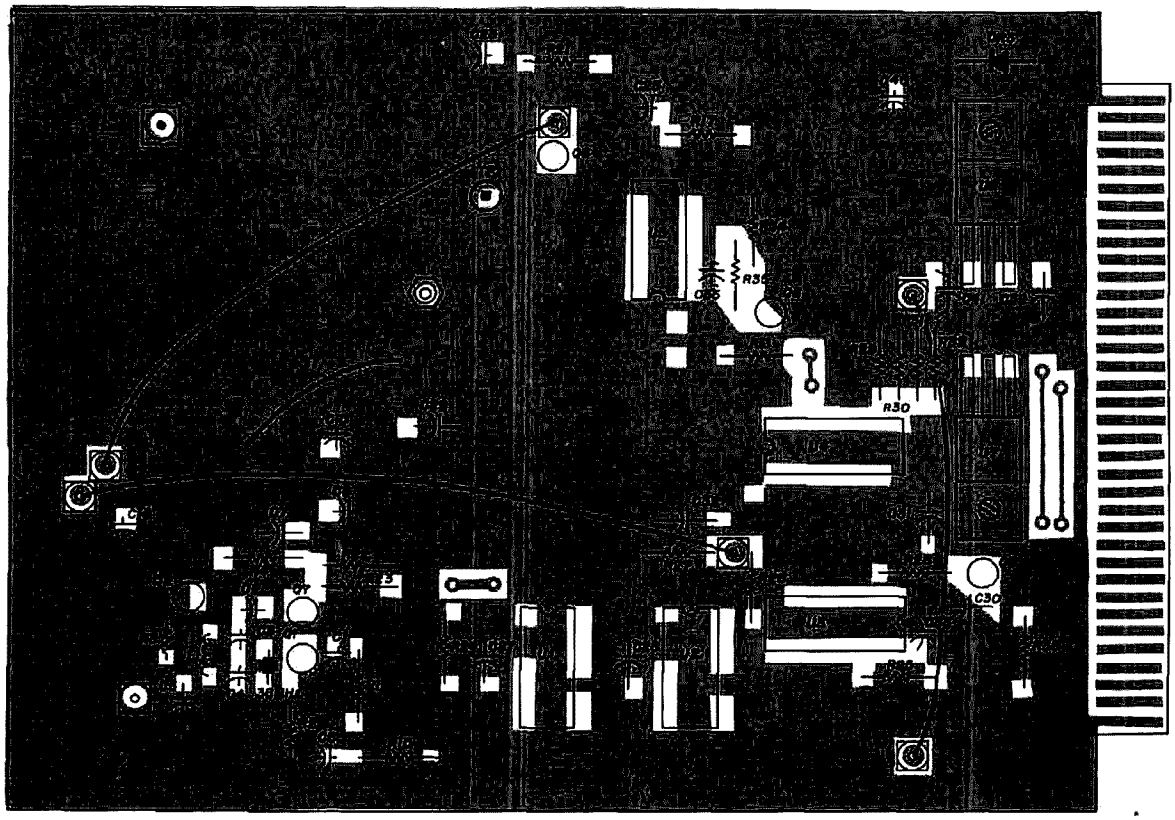


fig. 10(A). Component layout superimposed on groundplane artwork.

reach 4 volts, but as little as 2 volts are needed to drive U1.

- Connect a DVM or 50,000 ohm/volt meter to TP1. Tune the piston trimmer and note the meter response, which should be similar to the discriminator curve shown in **fig. 6**. The oscillator will be phase-locked if the voltage at TP1 falls on the sloping part of the curve. If the oscillator isn't operating, TP1 will be less than 2.0 volts, and possibly almost zero. If the voltage on TP1 is greater than about 7 VDC, the AFC (U3) is inoperative.
- Pins 3 and 14 of U3 should show square waves at the same frequency; the leading edges should be approximately in phase if the oscillator is phase-locked. You can check this with a dual-trace scope with a common time base. The square waves should have periods of 1.777 μsec.
- Adjust the piston trimmer until the square waves are exactly in phase. The voltage on TP1 now corresponds to the optimum operating point where phase noise should be at its lowest. On the prototype PC board, the optimum value was 4.0 volts DC.

an interesting experiment

To illustrate the increase in phase noise caused by large values of N, you can perform this experiment if you have access to a spectrum analyzer. Carefully solder a short piece of wire across the phase detector from the loop, leaving the CD4046 to perform both AFC and phase detection. Lockup will occur as before, and voltage levels at TP1 will be unchanged. However, the level of phase noise as seen on the spectrum analyzer will have increased by about 60 dB in and around the carrier and out to several MHz. Except for the fact that the carrier is stable, the spectrum appears only slightly cleaner than when the oscillator is free-running. This agrees fairly well with the phase-locked N2048 curve of **fig. 3** of part 1. **Figure 7** is an oscillogram of the VCO spectrum under these conditions, while **fig. 8.** shows the phase-locked spectrum.

I've chosen to call the PLL described here a "low noise" system. The title, of course, is relative, so further explanation is needed.

The experts all agree that measuring phase noise accurately is difficult at best and strewn with potential pitfalls depending on the technique used. Basically, phase noise level is defined as:

$$\frac{E_{sb}}{E_c} = 20 \log \left[\frac{\Delta f_{rms}}{\sqrt{2}\, f_m} \right] dB \qquad (1)$$

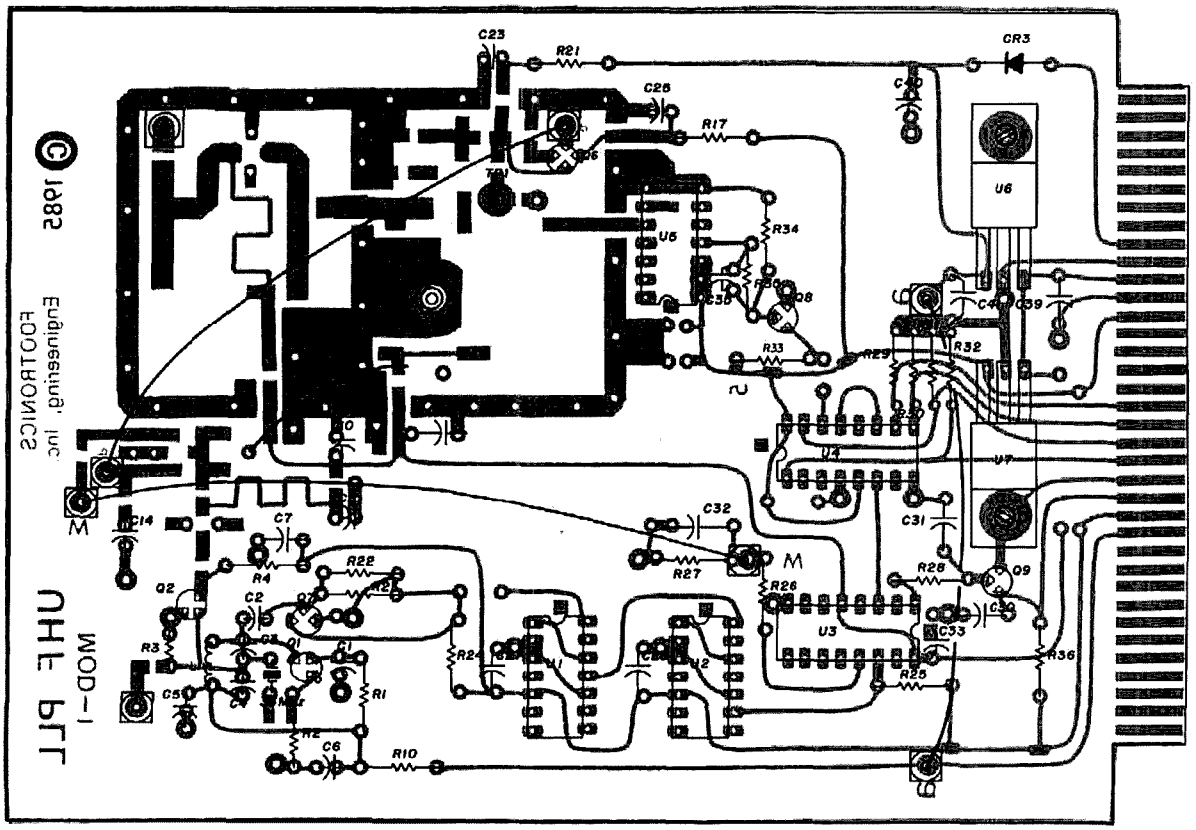


fig. 10(B). Component layout superimposed on trace side artwork.

where E_{sb} and E_c are respectively sideband and carrier voltages. Here, Δf_{rms} is the RMS value of peak frequency deviation at a distance f_m from the carrier. But how does one measure Δf? In high-purity sources, $20 \log (\Delta f/f_m$ is an extremely small number, typically more than 60 dB below the carrier. At an offset of 10 kHz, phase deviation of a phase-locked source may well be much less than ±0.001 degree.

In the absence of sophisticated test equipment, which is well beyond the reach of most Amateurs, what can an inexpensive analyzer such as the HP-8551 show us? One sees the RMS sum of all the power passing through the 1-kHz bandwidth. This can be extrapolated to a 1-Hz band by use of so-called stochastic or random theory:

$$\frac{W_{3000}}{W_1} \pm 10 \log \frac{B_{1000}}{B_1} \pm 30\ dB$$

In other words, the ultimate sensitivity of the 8551 is −(70+30) or −100 dBc/Hz. Based on this estimate, it should be easy to see the noise of free-running UHF oscillators. Judging from the spectrum response shown in **fig. 8**, the phase noise of the UHF VCO described here is less than −100 dBc/Hz close-in — considerably better than the calculated level of −85

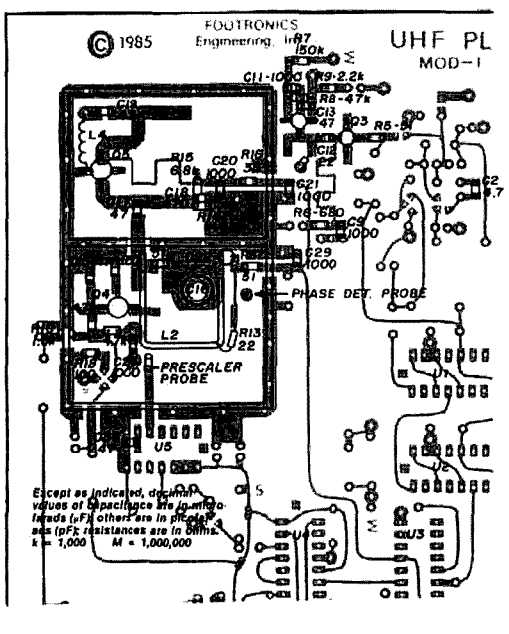
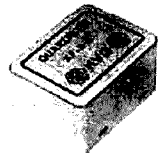


fig. 11. Detail of trace side of PCB showing locations of chip capacitors and resistors as well as other VCO buffer and loop amplifier components.

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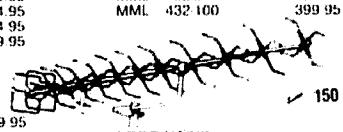
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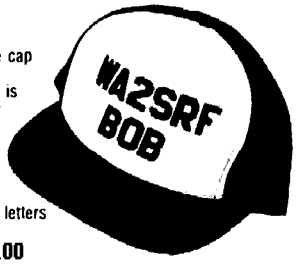
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dBc/Hz. At offsets beyond 10 kHz, the noise is below the resolving power of the spectrum analyzer.

conclusion

In this article, I've described a phase-locked UHF VCO for Amateurs and experimenters who need a very clean signal source. The performance of this VCO has been compared with an earlier type of signal source using a crystal oscillator multiplier scheme. Both models are shown to have excellent phase-noise characteristics, based on measurements made on a spectrum analyzer. The UHF VCO's level of coherent sidebands is over 60 dB below the carrier, a considerable improvement over the 1969 model. This characteristic is particularly important when the VCO is used to drive a multiplier for microwave applications. The VCO and associated circuits are mounted on a standard plug-in PC card. This assembly is typically dedicated to provide approximately +10 dBm of clean local oscillator power for UHF up- or down-converters. The VCO is also well suited for block down-conversion use.

The plated-through nature of the PC board is essential to good UHF performance. The complete VHF/PLL, double-sided printed circuit board artwork is illustrated in **fig. 9A** (ground plane side) and **fig. 9B** (trace side). The ground plane-mounted component layout includes two views in **fig. 10** — with the ground plane side **fig. 10A** and the trace side of the PC board **fig. 10B** provided in the background. Finally, a section of the trace side of the PCB shows the chip capacitors' and resistors' component layout superimposed on it in **fig. 11**. If you need help in obtaining this board or other components such as shield boxes, coils, diodes, transistors, and the like, send a business-sized SASE to me at the address shown at the beginning of the article.

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ham radio TECHNIQUES

Bill Orr W6SAI

50th anniversary of the tube that changed everything

1936 was a very good year. The country seemed to be coming out of the Great Depression. More people were at work and the lines in front of the soup kitchens seemed to be getting shorter . . . and great things were stirring in the young field of electronics.

For years circuit designers and radio manufacturers had struggled with low-gain, low-performance receiving tubes that consumed a lot of power and produced only mediocre results. The battery-operated receiver was dead; the AC receiver was here to stay. The time was ripe for the introduction of a new family of tubes that would transform radio from a curiosity into a piece of household furniture.

In 1935 the General Electric Company developed the idea of a metal receiving tube that wouldn't require an external shield and could be made in large quantity at low cost. Experimental tubes were made and the Radio Corporation of America was licensed to produce such tubes for the home entertainment market.

At the same time, Otto Schade, Sr., and his associates at RCA were at work on a revolutionary tube design that would offer high gain, high power output, and low signal distortion. It seemed a good idea to incorporate this concept in the newly developed metal envelope. The new tetrode tube made use of aligned grid and screen elements and beam-confining plates. This resulted in an electronic suppressor within the tube structure that increased overall efficiency and removed some of the "kinks" in the tube characteristics that had plagued older, less sophisticated designs.

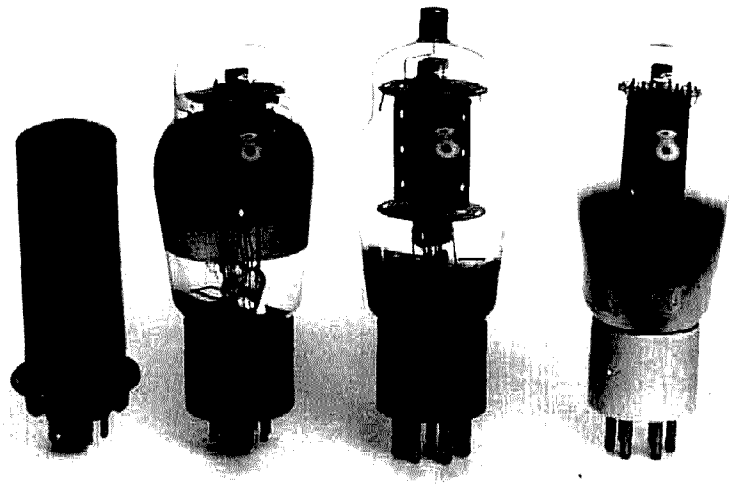


fig. 1. The famous 6L6 passed through several designs in its long lifespan. At left is the original metal 6L6. Next to it is the glass 6L6G version. Second from right is the transmitting version, the 807 (6.3 volts) or 1625 (12.6 volts). The 807 had two getters in the base of the tube, shield plates around the tube stem and extra insulation between the metal anode and the mica supports. The plate lead was brought out the top of the envelope. At right is the T-21, a transmitting version of the 6L6G. It employed a 6-pin ceramic base of low loss, but eliminated the anode insulating supports. The T-21 performed well at 500 volts but often flashed over when run at higher voltages. Even so, it was a bargain for less than two dollars.

The happy results of this pioneering work were announced in the early spring of 1936 with the introduction of the 6L6 "Beam Power Tube" (**fig. 1**).

too good to be true?

The characteristics of the new tube seemed too much to believe. A pair of 6L6s working with 400 volts on the

plate could produce over 25 watts of audio power! And even more interesting, the 6L6 could grind out over 35 watts as a crystal oscillator!

Now the big question — how would the 6L6 act as an RF amplifier? The tube was tried out; results were mixed. What to do with the metal envelope? Ground it? Let it float? There was much argument about this. *QST* wasn't sure what to do, so they left the shield floating in a simple amplifier design and remarked that — unfortunately — the tube had to be neutralized to operate stably.

The problem resolved itself later that year, when RCA and others announced a glass version of the 6L6 known as the 6L6G. This repackaging job was presumably done to reduce manufacturing costs and speed up production; it's thought that the metal tube manufacturing technique was not proving cost-effective. At the same time, the 6L6G was rearranged to bring the plate connection out of the top of the envelope. Thus was the 807 born.

In order to make the 807 a true transmitting tube, additional internal insulation was added to boost the maximum voltage rating to 600 volts and small shields were added around the base of the tube stem to reduce internal feedback paths. In reviewing the tube, *QST* said, "It may have a tendency to self-oscillate in high-gain circuits above 14 Megacycles, or in designs where input and output circuits [are] not carefully isolated."

life and death of the 807

The 807 was a blockbuster. It was the first high-gain, low-drive, inexpensive transmitting tube for the Radio Amateur. A pair of 807s — at $3 each — could comfortably run at 120 to 150 watts input on CW all day long!

Unfortunately, everybody got into the act. Clones of the 807 were made by many manufacturers, and cheaper versions of the 6L6G were sold with low-loss bases for RF service. Oldtimers remember the Taylor T-21, which sold for $1.95 and seemed to work as well as the 807.

But the 807 was to come upon rocky days. When it went into large-scale production during World War II, the special insulators, as well as the base stem shields within the tube, were eliminated to cut costs. The 807 began to look more and more like a simple version of the 6L6G (**fig. 2**).

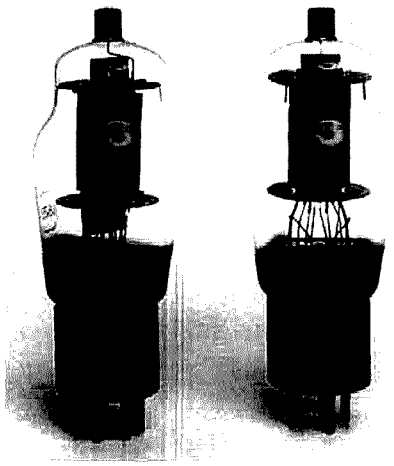


fig. 2. The pre-war 807 (1625). Note the base shield and the ceramic insulators between anode and mica supporting structure. At right is post-war 807. The base shield is missing and the ceramic anode supports have been removed. The changed electrical characteristics of the tube, plus the fact that many manufacturers were making the tube to slightly different specifications, tended to give the tube a reputation for instability and poor performance

The problem was that circuits designed for the "good" 807 couldn't perform properly with the cheaper version. The 807 rapidly earned a reputation as an unstable tube, mainly because so many different versions were available. In those days, manufacturers didn't know — or simply ignored the fact — that changing the internal support structure and moving the connecting leads to the elements about to fit a particular manufacturing process would drastically alter the operation of the tube, particularly at higher frequencies.

While the 807 faded into obscurity and was replaced by the newer 6146, the 6L6G flourished in the brand-new field of television. It was revived for hi-fi audio as the 6L6GC/6L6GB and various improved models saw duty as sweep tubes, with the final versions of the venerable design being the 6LQ6 and the 6MJ6. Some of these tubes saw duty as linear amplifiers in Amateur service (**fig. 3**).

All in all, the original 6L6 concept endured for 50 years in one form or another, and the principle of electron beam direction has been carried into the modern transmitting tubes. The 8877, 4CX5000A, and many other modern tubes are descendents of the original design worked out by Schade and his colleagues over 50 years ago. Hats off to these pioneers!

the horizontal square loop antenna

A letter arrived from Bob Morrison, AG9C, which may be of general interest:

The November, 1985, QST contained an article on a horizontal square loop antenna. Such strong claims were made that I decided to compare a full-wave horizontal square loop with a half-wave dipole using the numerical electromagnetics code computer program, NEC-2.

A perfectly conducting ground plane is assumed. Far-field radiation patterns in either horizontal or vertical planes show little difference between the antennas.

This is not surprising since both antennas consist of elemental horizontal dipoles. However, nulls in the zenith lobe occur at different antenna heights. Also, the loop is resonant at all harmonics, while the dipole is resonant only at odd harmonics.

This is not to say the horizontal loop is not a good antenna—it is. Just don't expect miracles.

the K4EF antenna and tuner

The "All-band" wire antenna and tuner designed by Ev Brown, K4EF, perform well on all bands between 160 and 10 meters (**fig. 4**). The antenna consists of a 200-foot wire, with 50 feet in the vertical plane and the remainder run horizontally to a nearby tiepoint. The antenna works against

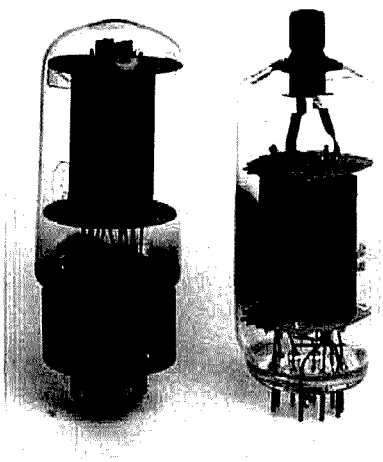


fig. 3. The ultimate 6L6. At left is the 6L6-GB with high plate dissipation rating (30 watts) and improved base structure. A single tube would deliver 11 watts of audio at 350 volts plate potential. At right is the sweep tube version of the 6L6—the 6LQ6. Note the heavy anode leads. Tubes of this type were popular in Amateur linear amplifiers for a number of years.

ground in Marconi fashion. Ev originally used a single ground rod, but later laid out a ground screen beneath the antenna consisting of 1000 feet of No. 14 wire forming a mat about 20 feet wide and parallel to the horizontal portion of the overhead wire. After all this effort, he could detect no change in performance!

One reason efficiency is high is that the feedpoint resistance of the antenna is about 100 ohms at 3.5 MHz and nearly 4000 ohms at 1.8 MHz.

The feedpoint is very "hot" on the 160- and 80-meter bands and a high voltage, transmitting-type capacitor is required to prevent flashover. Toroidal coils, wound on several stacked Amidon T-200 forms, were tried but proved unsatisfactory because of heating and inductance drift. The solution was air-wound inductors.

Generally speaking, the tuner will tune any length of wire over 40 feet long at 7 MHz or higher, provided the wire is not a quarter wavelength long. The antenna feedpoint resistance must be higher than 50 ohms in order to allow the tuner to do its job.

If the antenna is used exclusively on 160 meters, the antenna length can be chosen to match available components. For example, if the tuner capacitor is only 200 pF, antenna length should be increased to about 210 feet. The coil inductance must be increased to about 80 turns (40μH). Conversely, if a 1000-pF capacitor is available and you want to reduce the length of the antenna, a length of 175 feet is appropriate.

Ev says this antenna is outstanding on 160, 80, and 40 meters because the current loop is quite high in the air. On the higher bands he's able to compete with the "big boys" and frequently raises DX stations on the first or second call.

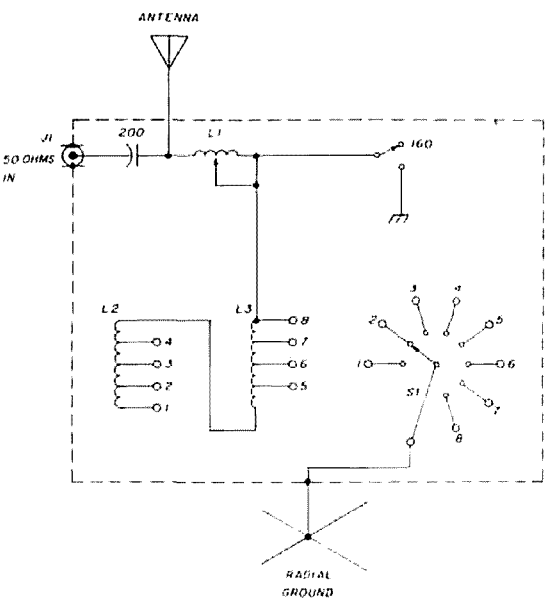
Ev likes the antenna so well he never got around to erecting his "Christmas tree" stack of Yagi beams on his 50-foot tower!

K1LPS on radials

One of the top DXers on 160 meters is K1LPS. Larry advises against being fooled by claims stating that a good ground (radial) system isn't required on 160 meters. "While a station using a vertical without radials will work stuff," he says, "a station using the same vertical with a good radial system will consistently outperform the equivalent antenna without radials on a given path."

K1LPS started out with an "L" antenna about 45 feet high with the end 90 feet running horizontally — quite a typical 160-meter antenna for the lean purse. He began with a single ground rod, plus four radials, none of which were more than 60 feet long, but found that results improved markedly as he improved his ground system. "The antenna was certainly not a pileup-beater by any means, but I eventually got there," he comments.

Encouraged by these results, Larry put up a better antenna. It was an L-shape, with 66 feet vertical and the remainder horizontal. He made careful measurements as he laid in a ground system, bit by bit. Starting out with a ground rod and four quarter-wave radi-



L1 Roller inductor, 35 turns, 2 1/2 inches diameter
L2,L3 B&W3900, 32 turns each, 2 inch diameter. Eight turns per inch. Tap every six turns.

Note: coax from station to J1 receptacle is wound into RF choke near J1, 7 turns, 18 inches diameter.

fig. 4. The K4EF wire tuner.

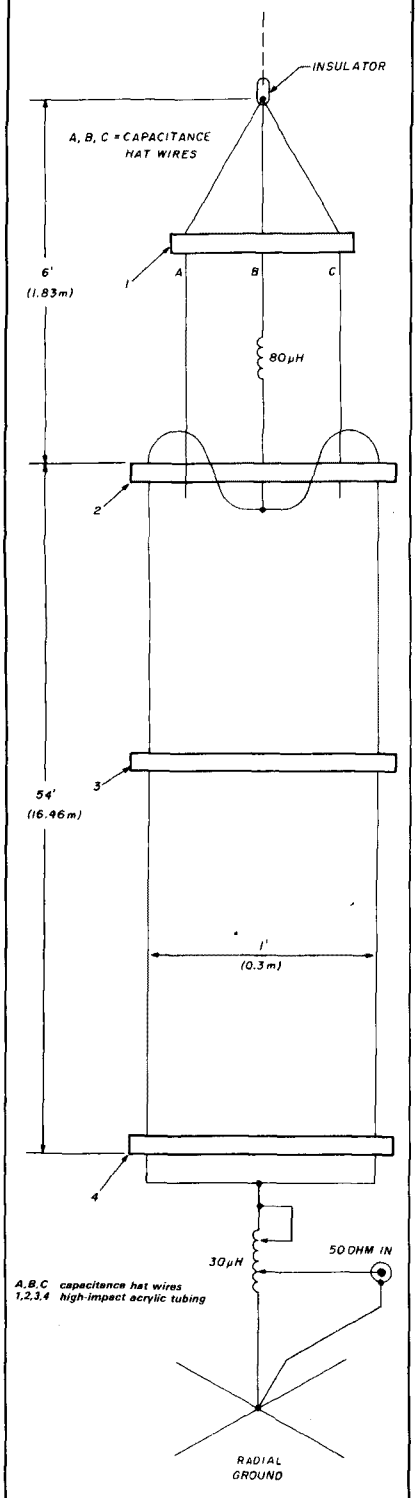


fig. 5. The W1FB 160 meter monopole antenna.

als, he measured his feedpoint resistance as 45 ohms. Bandwidth between the 1.5-to-1 SWR points was about 70 kHz.

He next added 18 additional quarter-wave radials. The feedpoint resistance dropped to 38 ohms and the 1.5-to-1 passband disappeared, because the minimum value of SWR at antenna resonance was 1.6-to-1! Before he built a matching network, he added ten more radials, bringing the feedpoint resistance down to about 22 ohms. Using a fixed matching network, K1LPS achieved a match of unity at 1820 kHz, but the antenna bandwidth between the 1.5-to-1 SWR points had dropped to 40 kHz.

These various changes took place over an extended period of time. Signal reports were solicited from many stations at various distances each night after additions to the ground system. A good many hams kept track of Larry's progress and there was a noticeable improvement as radials were added. It was finally concluded, however, that additional radials over 44 didn't seem to make much difference. And — finally — K1LPS made that elusive Japanese contact!

K1LPS says, "Yes, you can get by without radials. But if you want to work long-haul DX, find a way to get a good ground system installed. That's the difference between a passable signal and a pile-up breaker!"

In closing, Larry remarks, "A poor ground system has one thing going for it — a nice, broad SWR curve across a relatively wide frequency range. But then, so does a dummy load!"

the W1FB 160-meter monopole

Because it's difficult to be loud on 160 meters when you live on a city lot, I've long envied those lucky hams who live on country estates, with acres of land at their disposal. My good friend Doug, W1FB, has pondered the problem and finally come up with a compact monopole antenna design that solves this problem for Amateurs who can find 60 feet of space, either straight up, or at an angle (**fig. 5**).

Basically, the antenna is a "fat" monopole, top-loaded with an inductor and a capacitance hat. An arrangement such as this could end up being a rat's nest, but Doug has engineered the design through several models and arrived at a clean, compact, rugged package that's easy to install and adjust. Eighteen radials, each 130 feet long, are laid out beneath the antenna. The spacers are made of high-impact acrylic tubing and the array is made of No. 14 wire. Slung up in the clear, the measured SWR of the antenna between the 2-to-1 SWR points is 165 kHz. That's not too shabby for such a compact antenna!

More information on this nifty antenna? Write Oak Hills Research, 4061 North Douglas Road, Luther, Michigan 49656. Doug says the only problem now is that he doesn't have time to get on the air and work DX anymore!

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PRACTICALLY SPEAKING ... Joe Carr K4IPV

solid-state rectifiers

Because the lowly rectifier diode isn't exactly a high-tech electronic part, it's sometimes overlooked by those accustomed to more sophisticated electronics — microprocessors, packet radio TNCs, and HF SSB transceivers, for example. Nonetheless, the rectifier diode is extremely important.

Consider this: about 20 years ago there was a popular SSB rig on the market that suffered a terrible reputation for poor reliability. These rigs were on the blink more often than not (or so it seemed to those of us who owned them). The main — and for the most part *only* — problem was with the rectifier diodes. The manufacturer used a 1000-volt PIV rectifier in a 750-volt DC power supply. So what's wrong with using a 1000-volt diode in a 750-volt circuit? Read on, and you'll find out.

what are rectifiers?

The word "rectify" means "to make right or remove impurities." The purpose of a rectifier in a DC power supply circuit is to remove the impurities of the AC line current and make it right for DC-craving electronic circuits. AC is incompatible with most electronic circuits, so it must be changed to DC by a rectifier. From the above discussion you can probably guess that the main requirement for a rectifier is to convert bidirectional AC into a unidirectional form of current. Although ham equipment once used vacuum and mercury vapor rectifier tubes to accomplish this task, all modern equipment now relies on solid-state PN junction rectifiers.

Figure 1 shows the standard circuit symbol for the solid-state rectifer diode (**fig. 1A**), along with some common

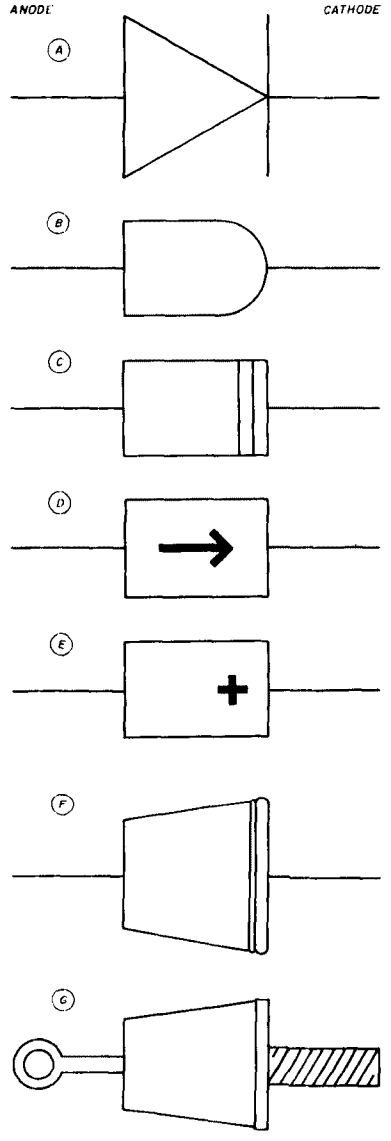


fig. 1. Solid state rectifier diodes: (A) standard circuit symbol; (B) through (E) common shape of epoxy or "plastic" package devices; (F) "Top-hat" type and (G) stud mounted.

shapes of actual diodes. The "input" side, where AC is applied, is the anode, while the "DC" output is at the cathode. The diodes shown in **figs. 1B through 1G** are positioned so that the respective anodes and cathodes are aligned with those of the circuit symbol in **fig. 1A**. Rectifiers in **figs. 1B** through **1E** are epoxy or "plastic" package devices, the type we usually see. The cathode end will be marked with either a rounded end **fig. 1B**, a line (**fig. 1C**), a diode arrow (**fig. 1D**) or a plus sign (**fig. 1E**).

The diode shown in **fig. 1F** is the old-fashioned — and now obsolete — "top-hat" type. Unless otherwise specified, the top-hat type can safely pass a current of 400 milliamperes, while those in **figs. 1B through 1E** generally pass 1 ampere (or more, for larger packages).

The stud-mounted type shown in **fig. 1G** is a high-current model. These diodes are rated at currents from 6 amperes or more; 50- and 100-ampere models are easily obtained. They are mounted using a threaded screw at one end, which also forms one electrical connection. The other electrical connection is the solder terminal at the other end. Unless otherwise specified, the solder terminal is the anode, while the stud-mount is the cathode terminal. Exceptions to the polarity rule are sometimes seen. The reverse polarity diodes will have either an arrow symbol pointing in the opposite direction (the arrow always points to the cathode), or an "R" suffix on the type number — for example, 1NxxxxR.

rectifier specifications

The use and abuse of solid-state rectifiers involve key specifications: for-

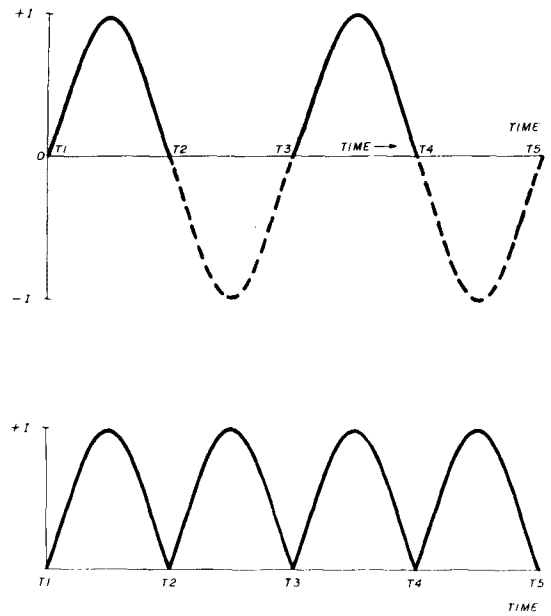


fig. 2. Rectifier output: (A) Solid line, half-wave rectified; (B) Full-wave rectified.

ward current, leakage current, surge current, junction temperature, and peak inverse voltage (PIV) — also called peak reverse voltage (PRV).

The forward current is the maximum constant current that the diode can pass without damage. For the popular 1N400x series of rectifiers the forward current rating is 1 ampere. The leakage current is the maximum current that will flow through a reverse biased junction. In an ideal diode, the leakage current is zero; in practical diodes of reasonable quality it is very low.

The maximum surge current is a rating that can get you into trouble. The surge current is typically very much larger than the forward current, so it's sometimes erroneously taken to be the operating current of the diode. It's defined as the maximum short-duration current that will not damage the diode. "Short duration" typically means 1/60 of a second! So don't use the surge current as if it were the forward current (a common but disastrous mistake).

The junction temperature is merely the maximum allowable temperature of the PN junction. The actual junction temperature depends on the forward current and how well the package (and environment) rids the diode of internal heat. Although typical maximum junction temperatures range from +125 to +150 degrees C, good design requires as low a temperature as possible. One reliability guide mentions that the junction temperature should be limited to a maximum of +110 degrees C.

The peak inverse voltage (PIV) is the maximum allowable reverse bias voltage that won't damage the diode. This rating is usually the limiting rating in certain power supply designs, and the least often heeded. Later in this article, we'll learn how to select the PIV rating for practical power supplies.

The output from the rectifier is not pure DC, but rather pulsating DC. For halfwave rectifiers we see a waveform such as **fig. 2A**, and for fullwave rectifiers we see the waveform of **fig. 2B**. The pulsating DC waveform is almost as useless for electronic circuits as the AC input. However, we can filter the pulsating DC output to form nearly pure DC. Although the subject of filtering is beyond the scope of this article, we must consider at least the simplest form of a power supply filter in order to correctly select a rectifier for any given application. **Figure 3A** shows a simple power supply with a brute-force filter capacitor, C, shunting the load to filter the pulsating DC into nearly pure DC. This capacitor will charge to a potential equal to the peak voltage, i.e. 1.414 times the applied RMS voltage.

selecting rectifier diodes

The two parameters that you'll most often use to specify practical power supply diodes are the forward current and peak inverse voltage. Get these right and in almost all cases the rectifier will work long and hard for you.

The forward current rating of the diodes must be at least equal to the maximum current load the power supply must deliver. That's common sense. But in the real world there's also the necessity for a safety margin to account for tolerances in the diodes and variations of the real load (as opposed to the calculated load). It's also true that making the rating of the diode larger than the load curent will greatly improve reliability. A good rule of thumb is to select a diode with a forward current rating of 1.5 to 2 times the calculated (or design goal) load current — or more if you can get it. Although selecting a diode with a very much larger forward current (e.g., 100 amperes for a 1-ampere circuit) is both wasteful and likely to make the diode not work like a diode, it's generally the rule to make the rating as high as feasible. The 1.5 to 2 times rule, however, should result in a reasonable margin of safety.

The peak inverse voltage (PIV) rating can be a little more complicated. In unfiltered, purely resistive circuits the PIV rating needs only to be greater than the maximum peak applied AC voltage (1.414 times RMS). If a 20-percent safety margin is desired, then make it 1.7 times RMS voltage.

Most rectifiers are used in filtered circuits (e.g. **fig. 3A**), and that makes the problem different. **Figure 3B** shows the simple capacitor filtered circuit redrawn to better illustrate the circuit action. Keep in mind that capacitor C is charged to the peak voltage

with the polarity shown. The voltage across the transformer secondary (V) is in series with the capacitor voltage. When voltage V is positive, the transformer voltage and capacitor voltage cancel out, making the diode reverse voltage nearly zero. But when the transformer voltage (V) is negative, the two negative voltages (V and Vc) add up to twice the peak voltage, or approximately 2.83 times the RMS voltage. Therefore, the absolute minimum value of PIV rating for the diode is 2.83 times the applied RMS. If you prefer a 20-percent safety margin (a good idea), then make the diode PIV rating 3.4 times RMS (or more).

The problem with the HF SSB rig referred to at the beginning of this column was that the PIV rating of the diodes was too low. The applied RMS voltage from the secondary of the power transformer in that unit was about 540 volts, with the peak voltage being approximately 750 volts. Thus, the maximum reverse voltage seen by those diodes was on the order of 1500 volts —- which is death for a 1000-volt PIV diode! The cure, which many Amateurs opted to use, was to replace each diode in the power supply with two diodes connected in series (see below).

using rectifier diodes

In most cases, especially in low-voltage DC power supplies used by many electronic circuits, we can get away with simply using the diodes as shown in the circuits above. In **fig. 4**, however, we see the so-called "proper" way to use the solid-state diode rectifier. The resistor in series with diode CR1 (i.e. R_S) is used to limit the forward current. Many circuits, especially those with capacitor-input filter circuits, exhibit a surge current at initial turn-on. This current can sometimes pop the diode, so R_S is used to limit the possible damage. The value of resistance used for R_S is typically 5 to 20 ohms, with a power rating of 2 watts. In most cases, however, we can eliminate R_S by using a diode with a rating larger than the load current (for example, the two-times rule).

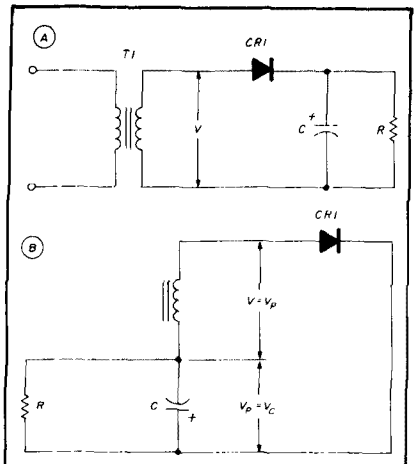


fig. 3. Simple half-wave rectification power supply; (A) Common circuit diagram; (B) Diagram redrawn to illustrate voltages present across each terminal.

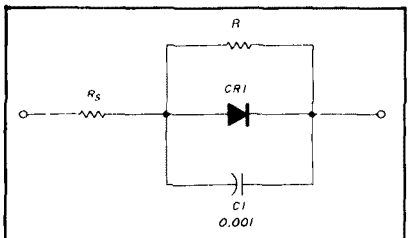


fig. 4. Proper way to use solid state rectifier. R_S limits the surge current experienced by the diode.

Capacitor C1 in **fig. 4** is used to bypass high voltage transient spikes around the diode. These spikes could possibly blow the diode PN junction. In fact, high voltage line spikes make diodes pop with regularity. Placing the capacitor in parallel with the diode will eliminate much of that problem. The working voltage (WVDC) of the capacitor should be equal to or greater than the PIV rating of the diode.

By using 1000-volt PIV diodes even in low-voltage circuits we can eliminate much of the damage caused by transients. We therefore would not need the capacitors. We can also eliminate the capacitors if a Metal Oxide Varistor (MOV) spike suppressor is used across the AC supply voltage.

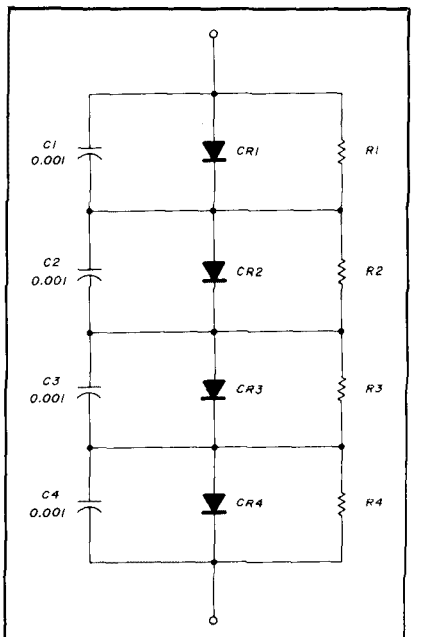


fig. 5. Method for using several diodes in series to increase the overall PIV rating.

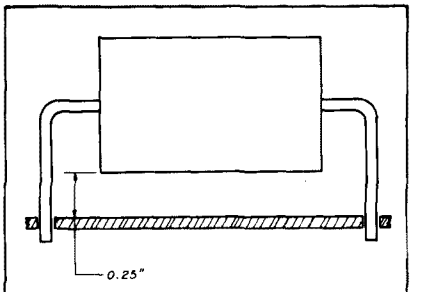
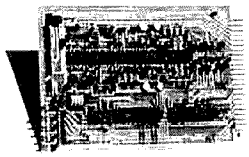


fig. 6. Preferred method of mounting an axial lead rectifier diode above a PCB to dissipate heat and prevent damage.

Figure 5 shows a method for using several diodes in series to increase the PIV rating. Assuming that the PIV ratings of the diodes are equal, then the overall rating is four times the rating of one diode. In general, if 1000-volt PIV diodes are used the PIV rating in the circuit is now 4000 volts. In the HF SSB transceiver we replaced the single diode with a pair of diodes connected exactly as shown in **fig. 5**.

The capacitors used in **fig. 5** are used for exactly the same purpose as those in **fig. 4**. The resistors, however,

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are needed for a different purpose. They equalize the forward voltage drop across each diode. A 470-kilohm, 1-watt resistor is typically used for 1000-volt PIV diodes. The wattage rating is required not because of the power dissipation of the resistors, but for the voltage rating — yes, resistors *do* have voltage ratings.

Figure 6 shows the proper method for mounting an axial lead rectifier on a perf or printed circuit board. This method is used anyplace where no excessive vibration is expected; this includes most sedentary projects or equipment. The space beneath the diode body allows air to circulate (keeping the diode cooler) and prevents diode heat from damaging the board.

conclusion

Although solid-state rectifiers are among the most common electronic components used by Amateurs, they are also among the most common causes of failure in homebrew projects, repaired equipment, and equipment that was poorly designed from the beginning. By following the rules given above, you'll be able to successfully select the correct rectifier — and prevent reliability problems that take you off the air.

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Simple modification allows HF and VHF packet operation with one TNC

using an RTTY terminal unit with the Heathkit HD 4040 TNC

When I first started running packet radio my primary mode was 2-meter FM. The TNC (terminal node controller) was already set up with the proper tone pair for 2-meter operation. I had the appropriate radio equipment, and there was a fair amount of operating activity in the area. Operating was also very easy: I just dialed in the correct frequency on the HT, set the volume, and was on the air.

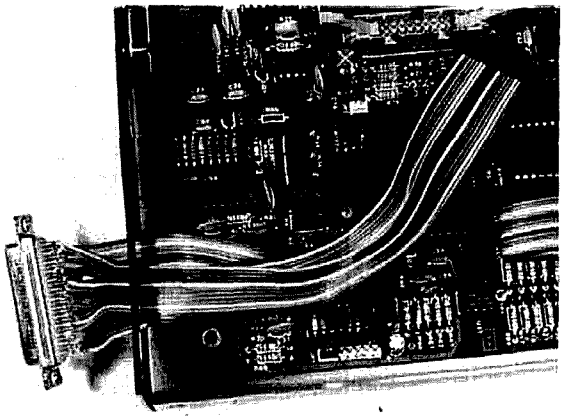
But soon I began to long for the excitement, challenge, and advantages of operating the HF Amateur bands. Packet was permitted on HF Amateur bands, but at a reduced baud rate and with a different tone pair than used on the 2-meter FM. It quickly became clear that to run both 2-meter and HF packet with the same TNC would require a bit of ingenuity.

the problem

When you first build the Heath TNC* it's set up for the tone pair used on VHF packet — 1200 Hz and 2200 Hz. The wide (1 kHz) shift allows for higher baud rates and lower error rates than are possible with a narrower shift. However, spectrum requirements on the HF Amateur bands require the use of a narrower shift, 200 Hz. The tones currently being used are 1600 Hz and 1800 Hz.

Though the on-board modem is very flexible, the difference in the tone pairs for the VHF and HF frequencies is too great for the modem to accommodate without changing the DIP headers and recalibrating the modem. (The DIP headers are the little cards with the resistors and capacitors on them that are plugged into the IC sockets.)

These changes, though easy to perform, do present a problem: every time you want to go from VHF to HF and vice-versa, you have to open the cabinet, change the headers, and recalibrate the modem. Although this isn't difficult, it's sufficiently troublesome to keep one from readily changing bands. In addition, the on-board modem lacks the type of filtering needed for reliable performance in the HF Amateur bands.



The wires have been soldered to the DB25 connector on one end and attached to a 20-pin header plug on the other. The header plug has been threaded through the hole in the back panel. The DB25 is ready to be mounted on the back panel and the header inserted into J5.

the solution

The designers of the Heath TNC included an on-board connector (J5) to allow you to use an

By Gregory P. Latta, AA8V, 438 Eastland Avenue, Akron, Ohio 44305

* The TAPR-1, AEA, and other clones are very similar to the Heath TNC. The modifications listed here should apply to them as well.

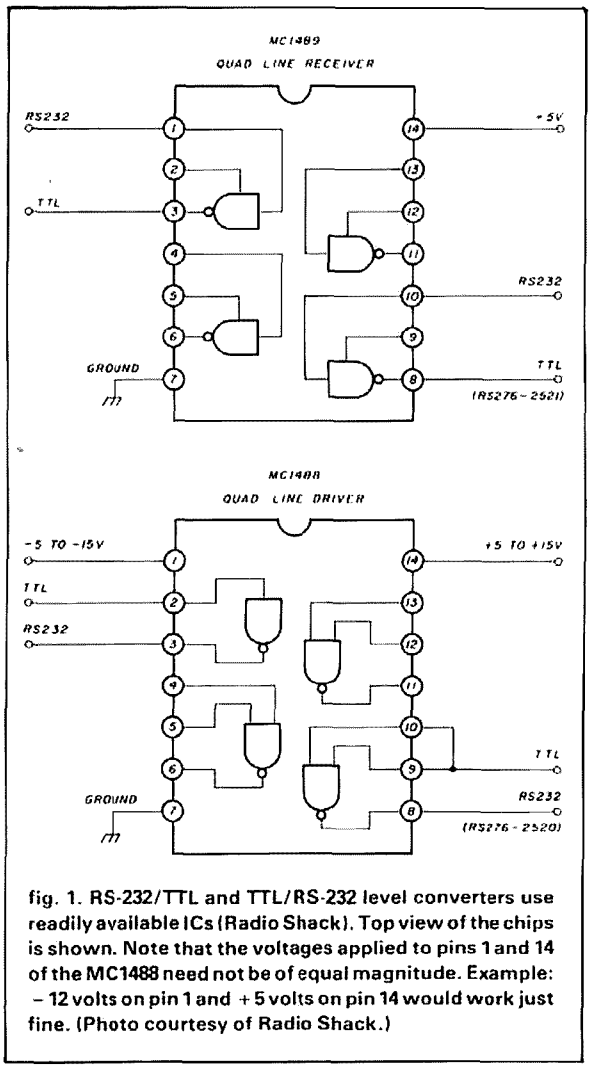


fig. 1. RS-232/TTL and TTL/RS-232 level converters use readily available ICs (Radio Shack). Top view of the chips is shown. Note that the voltages applied to pins 1 and 14 of the MC1488 need not be of equal magnitude. Example: – 12 volts on pin 1 and + 5 volts on pin 14 would work just fine. (Photo courtesy of Radio Shack.)

EXTERNAL modem (or TU) and bypass the on-board modem. (Thank you, Heathkit!) If the TNC were interfaced to an external terminal unit (TU), one would have the needed filtering, and changing bands would be reduced to switching a plug.

TU requirements

Certain minimum requirements that a modem (RTTY computer interface, terminal unit, etc.) must meet for operation with packet radio on the HF bands are detailed below:

•The TU must be able to operate at 200 or 170 Hz shift. (The difference between 200 Hz and 170 Hz is small enough that proper operation is not impaired.) *Note that the actual frequencies used are unimportant!* As long as the difference in frequencies (the shift) is 200 or 170 Hz, you can use the TU for packet radio. Just about all RTTY modems meet the 170 Hz requirement.

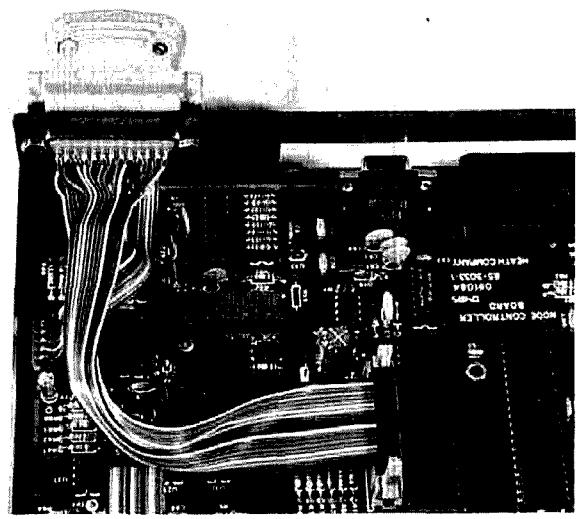
•The TU must be able to operate at 300 baud, the rate commonly used for HF packet operation. Some modems will not operate at 300 baud without modifications. Check the specifications of your TU to see if it meets this requirement.

•The TU *must* have some type of reliable tuning indicator. (If it doesn't you can devise your own.) A scope tuning indicator is fine. Because of the nature of packet radio, you cannot "tune until you get the print." You must be able to tune in the signal "blind" without looking for a print-out. If you can tune in the signal with your printer off, then flip on the printer and get good copy on RTTY, your tuning indicator is probably OK.

•The TU must have an output that indicates when the frequency is busy (i.e., when a Mark or Space is present). Packet radio is computer controlled and it is necessary for the TNC to be able to tell whether someone is using the frequency.

Some TUs have an RDA (Receive Data Available) output. This is ideal. If your TU lacks such an output, don't despair. Some have an autostart circuit that may be modified to provide the needed output. If your TU has no autostart circuit or RDA output, you'll have to devise your own carrier detect circuit. (See below.)

•The TU inputs and outputs *must be TTL compatible. RS-232 inputs and outputs will not work and may damage the TNC!* If in doubt, measure the voltages on the TU output pins under both on-off or mark-space conditions. If you ever obtain a negative voltage or a positive voltage greater than 5 volts, chances



Completed modifications to the TNC. The DB25 has been mounted on the back panel, and the 20-pin header plug has been plugged into J5. The jumper for VHF operation (as shown in *fig. 3*, but now with hood installed) has been plugged into the DB25 connector and can be seen at the top of the photo.

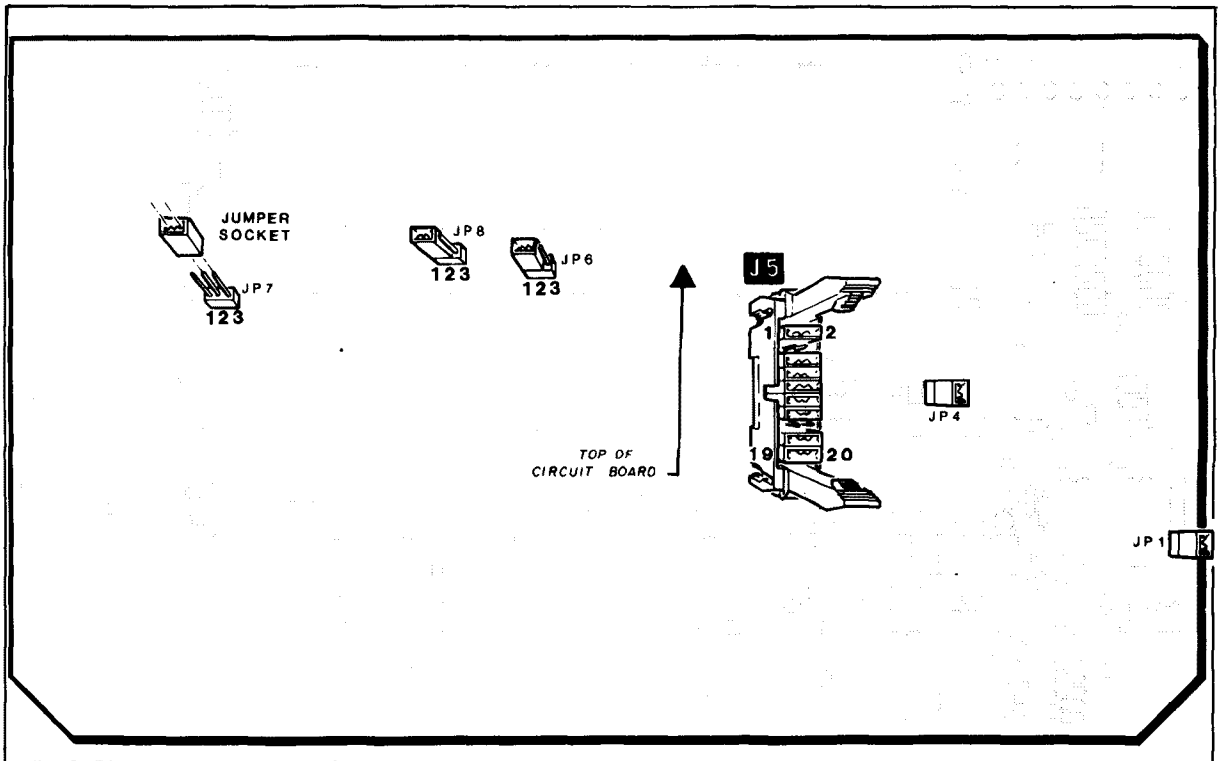


fig. 2. Diagram from the Heathkit manual shows the header plug J5 pinouts. The WD 1935 HDLC chip is to the left of J5, but is not shown in the figure (see *photo 5* as well). Pins 1-20 are connected to pins 1-20 respectively on the back panel mounted DB25 connector. (Photo courtesy of Heathkit.)

are that the output is RS-232 and should not be used. Some of the better units have both TTL and RS-232 inputs and outputs. Be sure to use those designated TTL.

If you have only RS-232 inputs and outputs you can build a simple circuit to convert them to TTL (see **fig. 1**).

TU candidate

I had been using the Heath HD-3030 RTTY Computer Interface on both VHF and HF RTTY for some time. This unit had a wonderful tuning indicator built in and the TU had proven its reliability under noisy conditions. It could also operate at 300 baud, the rate used for HF packet. In addition, the HD-3030 TU had an RDA (Receive Data Available) output that could be used as a "channel busy" indication. TTY inputs and outputs were also available. It met all the requirements mentioned above, and thus was an ideal choice for use as an external modem.

modifying the TNC

Before modifying the TNC, it is strongly recommended that you first get it working on 2-meter FM or some other VHF frequency. This way you can be sure that the TNC is operating properly. If you don't have a VHF radio, find a ham who's running packet radio and let him or her hook your TNC into his or her system. This will also allow you to gain some familiarity with the operation of the TNC. After you're sure that the TNC is working properly, you can start the modification.

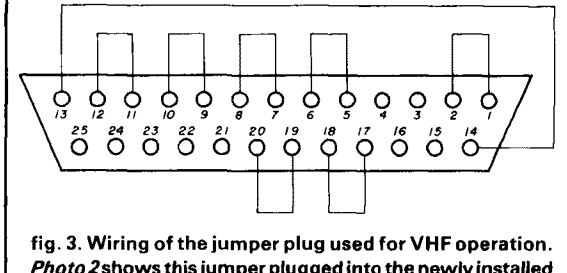


fig. 3. Wiring of the jumper plug used for VHF operation. *Photo 2* shows this jumper plugged into the newly installed DB25 connector.

You'll need the parts shown in **table 1.**

The first step is to make the hole for mounting the female DB25 connector on the back panel of the TNC in the area marked "caution." (Be sure to remove the board while making the hole for the connector.) Because it's much easier to solder the ribbon cable

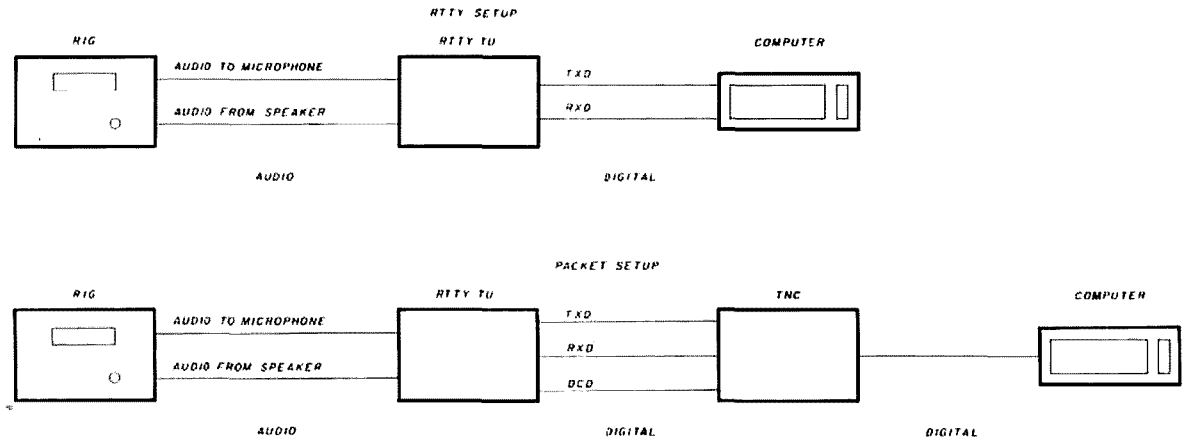
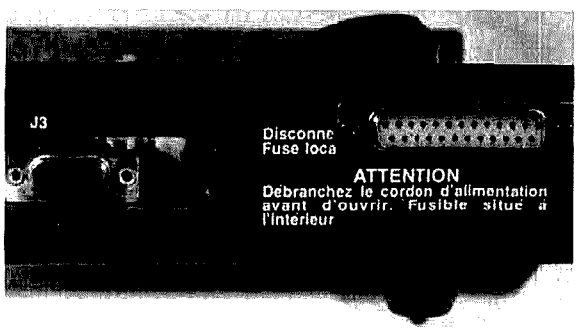


fig. 4. Comparison of RTTY and packet setups when an external modem is used for packet operation. Clarifies which signals are audio and which are digital. Note that the PTT lines have been omitted for clarity.



Rear view of the TNC after modifications have been completed. J3 is the transmitter jack. (See also *fig. 5.*)

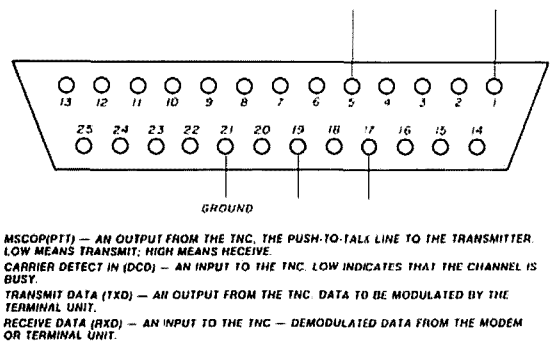


fig. 5. Diagram and explanation of the important pins on the newly mounted connector as viewed from the back panel. (See also *photo 3.*)

to the connector before it has been mounted, make the hole but don't mount the connector yet. After you've cut the appropriate mounting holes, re-install the board in the cabinet.

Next, connect the 20-pin header plug to one end of the ribbon cable. Be very careful to keep track of which pin is pin 1 and which is pin 20 (see **fig. 2**). Now solder the other end of the ribbon cable to the DB25 connector. The wire from pin 1 of the header plug should go to pin 1 of the DB25. The wire from pin 2 of the header plug should go to pin 2 of the DB25, and so on up to pin 20. *Be very careful not to get any of the wires crossed.* Check the cable with an ohmmeter when you're done. Finally, connect the solder lug to one end of a 3 inch (7.62 cm) wire. Connect the other end of the wire to pin 21 of the DB25 connector.

Feed the header plug, ribbon cable, and wire with solder lug through the mounting hole from the back and mount the DB25 connector (**photo 1**). Connect the solder lug to the closest available ground. (A circuit board mounting nut is perfect.) Plug the header plug into the header at J5, put on the cover, and you're all done! (See **photo 2.**)

So far, what we've done is to move the modem connector from the circuit board to the back panel. For

table 1. Parts needed for modification of the TNC.

- 25-pin female DB25 connector
- 25-pin male DB25 connector and hood
- No. 6 solder lug
- 20-pin connector to fit header J5
- about 9 inches (23 cm) of 20-conductor ribbon cable

normal operation, it will be necessary to make a jumper plug to replace the jumpers that were originally at J5. To make a jumper plug, take the male DB25 and solder jumper wires between the following pins: 1-2, 5-6, 7-8, 9-10, 11-12, 13-14, 17-18, 19-20 (see **fig. 3**). Put the hood on the connector and plug it into the previously installed DB25 connector.

That's it! At this point, you should verify that the TNC functions normally on VHF. If it does not, then there's a wiring error somewhere and it must be fixed.

between the TU and TNC

Even the most experienced ham is likely to be intimidated by the external modem section of the TNC manual, in which each pin ofJ5, the modem connector, is described in great detail. Fortunately, however, we need concern ourselves with only four of the 20 pins: two of these pins are used for input, and two are used for output from the HDLC chip. Before proceeding with the details, a description of these inputs and outputs is in order.

TNC digital inputs and outputs

When the TNC is operated with an external modem, it's important to realize that the inputs and outputs are not audio signals, but digital signals. It is the purpose of the external modem (MODulator-DEModulator) to convert the digital signals into an audio form that can be used by the radio link, and vice-versa. The only difference in using your TU for packet instead of RTTY is that instead of hooking it directly to your computer, you hook it to the TNC, and it's the TNC that communicates directly with your computer (see **fig. 4**).

There are four lines, besides ground (GND), that connect the TU to the TNC. Two go from the TNC to the TU, and two go from the TU to the TNC. The pin designations are for the DB25 connector installed earlier (see **fig. 5**). Recall that there's a 1:1 correspondence between pins 1-20 of J5 and pins 1-20 of our newly installed connector.

The two output lines that run from the TNC to the TU are labeled "MSCOT" and "TXD."

MSCOT (PTT) (Pin 5) is an output from the TNC to the modem. The name is, unfortunately, a real misnomer: MSCOT stands for "Miscellaneous Output," but this is nothing more than the TNC's PTT (Push-To-Talk) line. When the TNC drives this line low, the transmitter should be keyed or the TU should go into SEND mode. This should be connected to the TU's SEND or PTT input. *Do not use this pin to directly key your transmitter.* Doing so could damage the TNC.

TXD (Pin 19) stands for "Transmitted Data." The data output of the TNC, will normally be connected to the TU's "AFSK INPUT." (AFSK stands for "Audio Frequency Shift Keying.")

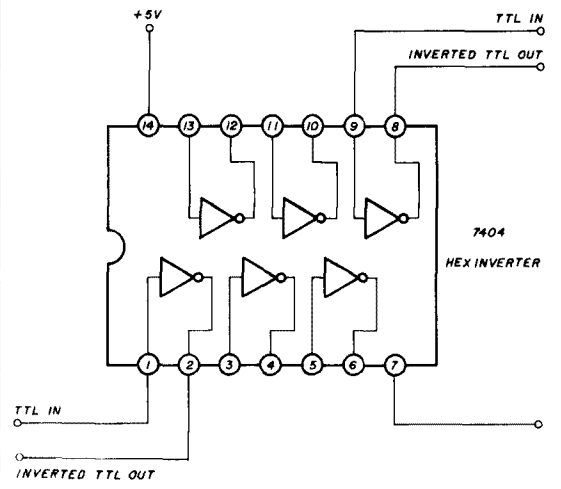


fig. 6. Circuit used for inverting TTL signals, necessary when a TTL signal on a non-Heath TU is inverted (i.e., when RDA is high when it should be low, etc.). The 7404 is available at Radio Shack (part No. 276-1802). Top view of the chip. (This circuit, reproduced with permission, originally appeared in Don Lancaster's *TTL Cookbook*, published by Howard Sams & Co., Inc., 1978.)

table 2. Connections for the HD-3030 RTTY TU.

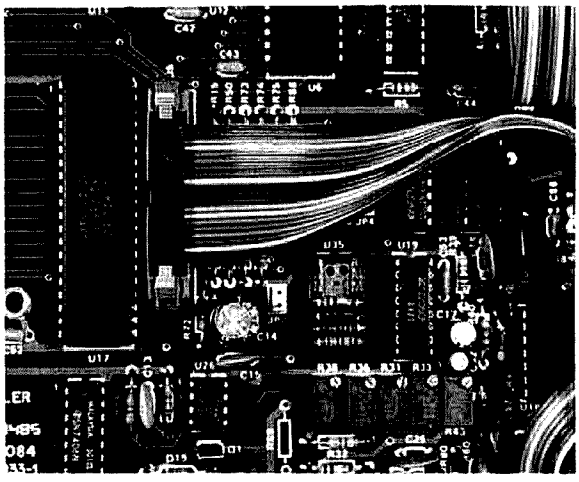
TNC pin number		Heath TU pin number
1 (DCD)	to	1 (RDA)
5 (PTT)	to	9 (SEND-N)
17 (RXD)	to	3 (DMOUT-TTL)
19 (TXD)	to	5 (AFSKIN-TTL)
21 (Ground)	to	21 (Ground)

The two input lines running from the TU to the TNC are labeled "Carrier Detect In" and "RXD."

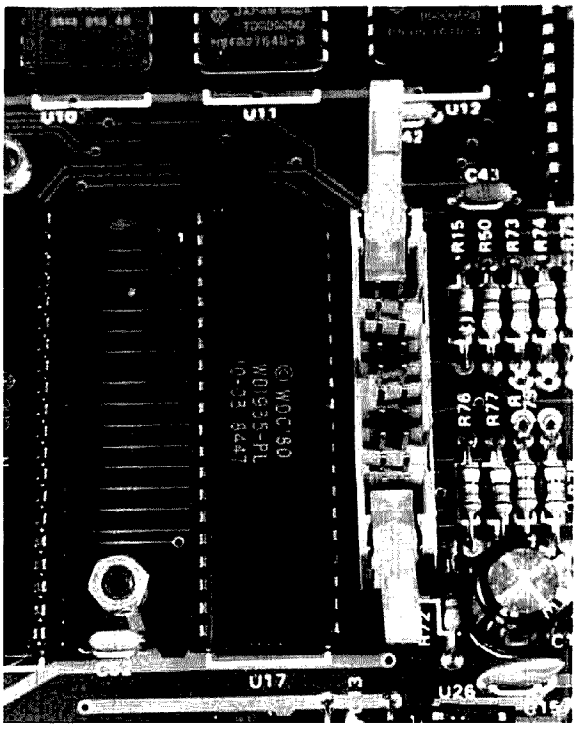
CARRIER DETECT IN (Pin 1) — or "DCD" for "Data Carrier Detect" — is an input to the TNC that tells the TNC when the channel is busy. A low on this line indicates that a carrier (mark or space) is present and that the TNC should not transmit. This is usually connected to the TU's "RDA" (Receive Data Available) output. Note that a low indicates that the channel is busy. If your TU's RDA output is high when the channel is busy, you'll have to use an inverter circuit (see **fig. 6**).

If you don't have an RDA output on your TU, you may be able to make use of its "autostart circuit" if it has one. What you need is a TTL output that's low (0 volts) when mark or space is present, and high (3-5 volts) when mark or space is NOT present. Autostart circuits vary greatly, so you're on your own here.

RXD (Pin 17) is the "Received Data" input to the



Closeup of J5 with 20-pin header plug installed. (See also *fig. 2*).



Closeup of J5. The WD-1935 HDLC clip is seen to the left of J5. (See also *fig. 2.*)

TNC. After your TU has decoded the audio tones and converted them back into a digital signal, that signal is applied to this input. This normally goes to your TU's "DMOUT" (DeModulator OUTput) pin.

These four lines, in addition to a ground lead, are all that's needed to connect the TNC to your TU!

By the way, don't worry if the data lines "TXD" and "RXD" might be inverted or "upside down." It's the TRANSISTIONS from mark to space that count in packet radio. The result is that the data can be "right side up" or "upside down." It will work either way!

HD-3030

The connections given below are for the HD-3030 RTTY TU. If you plan on using a different TU, you'll have to work out the connections based on the information given above.

You'll need two 25-pin DB25 male connectors and hoods and a length of 4-conductor (plus ground) cable long enough to run from the TU to the TNC. Keep it as short as possible. Shielded cable is preferred. Make the connections shown in **table 2.**

It will also be necessary to solder jumpers between the following pins on the TNC connector: 7-8, 9-10, 11-12, 13-14. Note: these jumpers should be installed no matter what TU you use (see **fig. 7**).

That's all there is to it, (although you must, of course, also make the appropriate connections to the transmitter and receiver as you did for RTTY operation). Use a label maker to label the connectors so you can keep track of which goes to the TNC and which to the TU.

modifications to the TU

The modifications to the Heath TU are minimal. The time constant of the RDA circuit is much too long for packet operation. To reduce the time constant to a value suitable for both RTTY and packet, replace capacitor C29 (a 4.7 μF electrolytic) with a 0.047 μF capacitor. If you don't plan to use the TU with RTTY, then a value of 0.01 μF is more suitable. Some readjustment of the RDA threshold may also be necessary, as explained below.

making it work

If you're still with me and have performed all the necessary modifications (there really aren't that many), you should be ready to start operating. Unplug the jumper from the back of the TNC and connect the previously prepared cable from the TU to the TNC. Make the appropriate connections between the TU and your radio, and between your computer and the TNC.

The various settings of the TNC can make or break the performance of your packet system. I recommend the following settings as a starting point:

PACLEN	60
HBAUD	300
AXD	6
DWAIT	0
TXD	1
FRACK	6
MAXFRAME	1

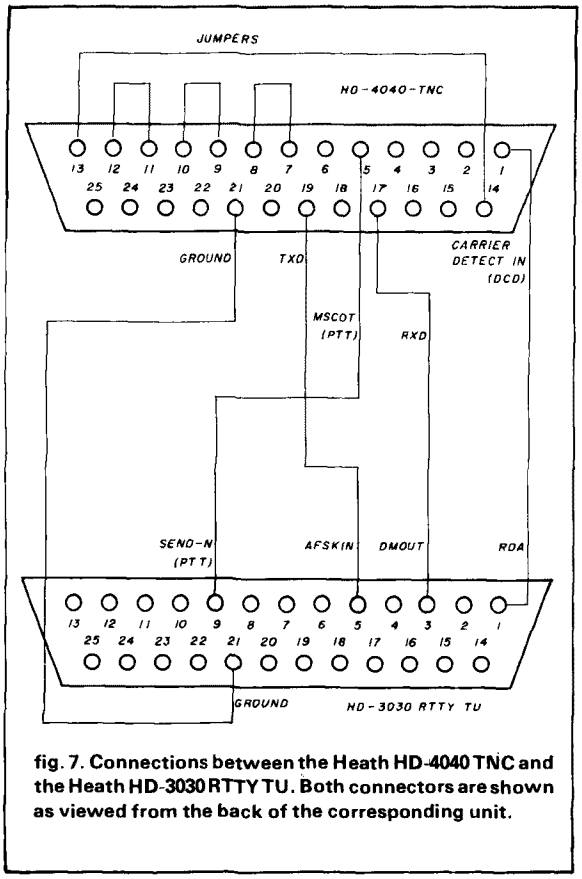


fig. 7. Connections between the Heath HD-4040 TNC and the Heath HD-3030 RTTY TU. Both connectors are shown as viewed from the back of the corresponding unit.

The time settings AXD, DWAIT, and TXD will give adequate key-up delay while minimizing the delay in transmitting once the channel clears. The parameters PACLEN, FRACK, and MAXFRAME are set to minimize the lengths of packets and transmitter "on" time. This results in fewer retries and increased throughput on the frequency. You can, of course, experiment with all of the settings except HBAUD, which must be set to 300.

NOTE: There's a bug in the resident TNC software (Version 3.2). The TNC won't properly receive packets at 300 baud until it has sent a packet or done a CW identification. Also, on powerup, you may get an "HDLC Can't Init" error. Don't worry. This is also easily solved by making the TNC identify. Therefore, after setting HBAUD to 300, or after powerup, enter ID <RETURN>. This will cause the TNC to identify and initialize the HDLC chip so that you can receive packets.

It's best to gain some listening experience before jumping into your first QSO. Tuning is critical, and there's no point trying to transmit packets unless you can receive them.

Make sure that you've enabled all of the TNC monitor functions: (M ON, MALL ON, MCON ON, MFROM ALL, MTO ALL, MAX FRAME 1). Next, set your receiver bandwidth to approximately 1 kHz. (After receiving packets you may be able to use a narrower bandpass, but in some receivers 400 Hz to 500 Hz is too narrow for proper operation.) Tune your receiver to one of the HF packet frequencies currently in use: 7.097, 14.103, 14.105, or 14.107 MHz. The latter is a good choice to start with because it's usually active at all times of the day, and signals are usually fairly strong.

Packet signals make a very distinctive whirring noise. But you must tune quickly, because individual packets don't last long. Tune as if you were tuning in an RTTY station, but note that the packet will not be displayed until all of it has been received. (This is why you must be able to tune "blind.") Also, if there are any errors in the packet, it won't be displayed. After a little practice you'll get the hang of it, and packets will be flying across your screen!

Transmitting is no problem, although you may find that the TNC waits a long time before transmitting the packet. These long waits (sometimes almost forever) are caused by the TNC's "thinking" that the channel is busy when it really isn't. This usually indicates that the signal on the "Carrier Detect" or DCD input is hanging on too long. With the Heath TU, this means that the RDA (Receive Data Available) indicator is hanging on too long. To keep this from happening, reduce the receive audio to the minimum needed to trip the RDA circuit. Also adjust the RDA threshold so that the circuit trips at the highest allowed audio level. These settings will cause the RDA light to go off as quickly as possible once the frequency is clear.

Other TUs may also exhibit the same problem. The cure is to simply find a way to reduce the time constant of the circuit driving the "Carrier Detect" or "DCD" input of the TNC to a suitably low value. If the time constant is too long, the TNC will think the frequency is always busy and never initiate a transmission.

conclusion

This TNC-RTTY TU combination has been in use for almost a year with good results. Switching between HF and VHF operation is very easy, since all one need do is remove the jumper on the back of the TNC and plug in the TU.

I hope that the information given here will enable many people to get in on the fun of HF packet radio. The interfacing information under "Digital Inputs and Outputs" should also be useful to those VHF operators who want to try using very high speed modems or some of the exotic modulation techniques that are permitted on the VHF bands.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

microwave and millimeter wave propagation: part 2

Last month's column, an introduction to microwave and millimeter-wave propagation, provided tables of frequency allocations and DX records established on these frequencies.[1] It stressed that at present, the primary mode of propagation on these frequencies is line-of-sight (LOS).

It was further noted that the apparent attenuation of an RF signal between two stations is more of a *physical* phenomenon than an *electrical* parameter. This is because of the decrease in the wavelength or size of a dipole antenna and hence its capture area. Therefore, if transmitter power, receiver sensitivity, and the size of the antenna are kept constant, the signal-to-noise ratio will actually *increase* as you go higher in frequency! This apparent "path loss" can be calculated using the equations and graph found in reference 1.

Ironically, the same propagation modes that serve as media for DX are the bane of commercial microwave users because they can induce path disruptions, outages, or interference. These modes — which microwave users often call "non-standard" or "anomalous" propagation — are the very ones that Amateurs are interested in exploiting for DX purposes.

tropospheric propagation

LOS propagation assumes that both stations can "see" each other. But this would be true only if you were operating in a complete vacuum, where the speed of light is constant. In the Earth's atmosphere, the presence of atmospheric gases and water molecules causes radio waves to travel more slowly.

In a standard atmosphere, the barometric pressure, temperature, and water vapor content all decrease linearly as altitude increases. Atmospheric density — the combination of these properties — also decreases with altitude. When microwave or millimeter-wave signals pass from dense air to thinner air, they undergo a change in direction proportional to the difference in density; this happens because the portion of the wavefront that enters the thinner air layer first begins to travel faster than the portion still in the dense air.

The result is a bending or refraction towards the denser air, which causes the signals to follow the contour of the Earth. Hence under normal weather conditions you can usually work about 25 to 50 percent further than the LOS path. If there's an extreme drop in atmospheric pressure with increasing altitude, anomalous or possibly extended propagation may occur; Amateurs often refer to this phenomenon as a "tropo" opening.

The amount of bending, known as the *refractive index gradient*, is more frequently referred to as the Earth's radius factor, or K. During standard atmospheric conditions, K will be 1.2 for dry elevated areas and 1.33 for inland areas. What this means is that under normal conditions, the typical path distance is 1.2 to 1.33 times the LOS distance.

However, K may reach 2 or 3 in humid coastal areas, which explains why coastal stations often enjoy improved propagation, especially over water. If an abnormal weather condition is present, the K index may go below 1 or even negative. If this happens, the path may become trapped or completely blocked, greatly attenuating the signals on a given path.

tropo bending and super-refraction

One of the most widely used DX propagation modes employed on the microwave and millimeter-wave frequencies is tropo bending. This often happens when the refractive index, K, goes positive beyond a value of 2.

If K gets very large, super-refraction takes place and microwave and millimeter-wave signals travel relatively unattenuated. They follow the curvature of the Earth for considerable distances until they reach a point where the K index drops back to normal. Methods of calculating the refractive index are discussed in references 2 and 3.

A rise in the K index results from changes in the atmosphere that result from abnormal weather conditions. For example, cool air passes over a warm body of water, causing evaporation of the water and a subsequent temperature inversion. Or a cold front may pass over an area of warm, stable weather, causing a marked decrease in total moisture content as altitude increases.

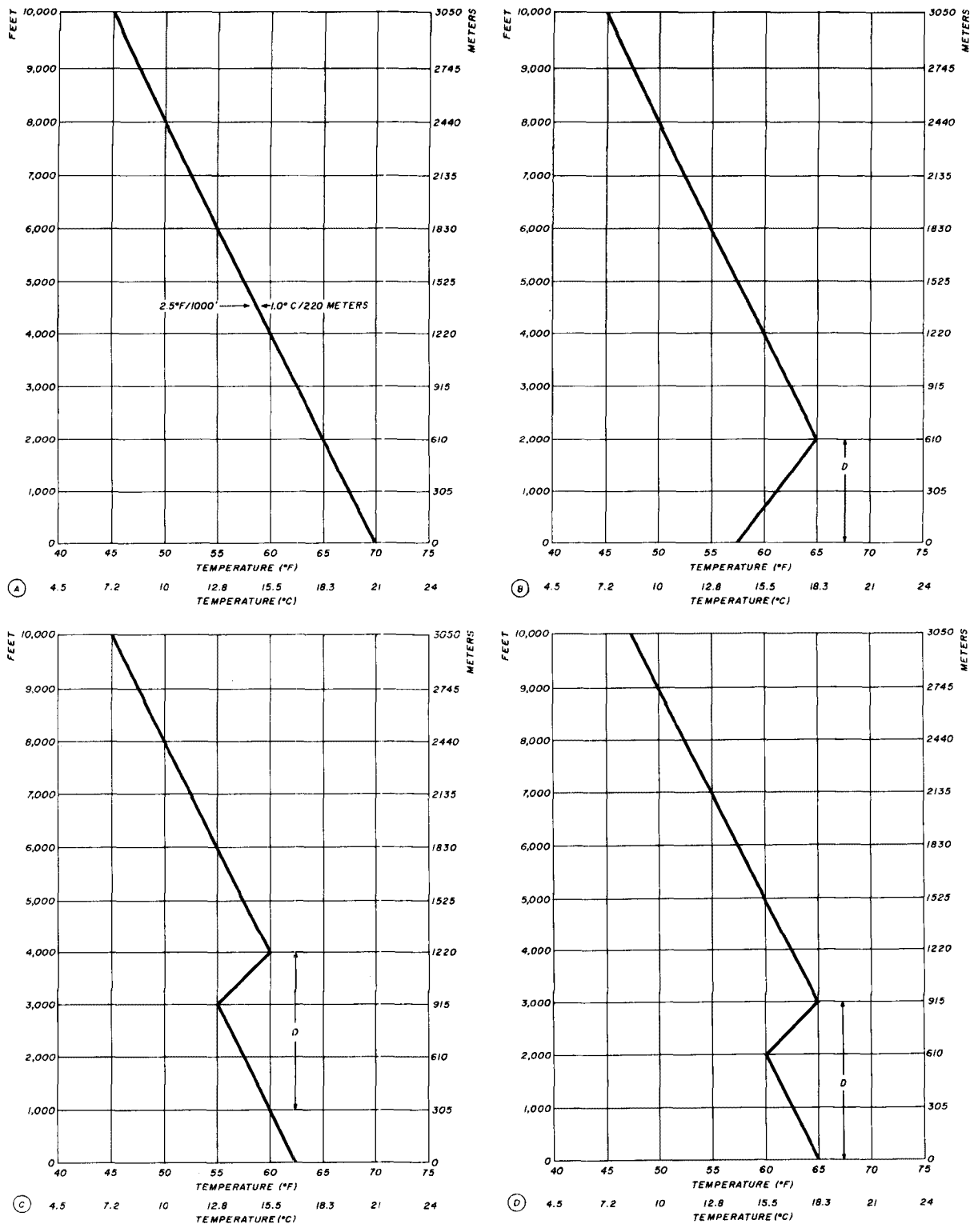


fig. 1. These graphs are examples of the standard atmosphere and typical ducts. (A) Temperature versus altitude graph of the "Standard Atmosphere." Temperature decreases approximately 2.5 degrees F per 1000 feet (1.0 degree C per 220 meters). (B) Typical evaporation duct that often happens over warm bodies of water is caused by a temperature inversion near the surface. "D" is the duct height. (C) Elevated duct. (D) Elevated duct that has its lower limits right at the surface.

These conditions typically occur between late spring and fall, but can happen at any time of the year. Openings also tend to be more pronounced in the evenings and around sunrise. They occur less frequently around noon, when the sun tends to warm the layers more uniformly.

In the southern portions of the continental United States, especially near the Gulf of Mexico, tropo openings begin in the late winter and continue through the spring. Optimum DX conditions across the Pacific between the west coast of the continental United States and Hawaii tend to peak in the summer. In the regions north and east of the Mississippi River, the longest overland DX via tropo bending tends to peak in the late summer and early fall.

At colder temperatures, the contribution of water vapor to refractivity is small because the saturated vapor pressure is low. However, at higher temperatures, humidity plays an increasingly important role in refraction. Therefore, refraction changes are least expected at northern latitudes but can be severe in hot, moist climates, especially in the equatorial regions.

The best DX conditions occur when the weather has been warm with high barometric pressure typically above 30.3 inches, and especially when a high pressure region is stalled or moving slowly across an area.[4,5] Then, if a weather front approaches, the high will move away, the barometric pressure will fall, and an opening may occur even if the barometer drops lower than 30 inches. I've noticed, by the way, that the best contacts take place between stations that have similar barometric pressure.

ducting

When the K index gets very high, a duct may form near the surface of the earth. These ducts act like a waveguide and propagate signals with extremely low attenuation. The height of the duct determines the range of frequencies that will propagate best within the duct.

Figure 1A shows the so-called "standard atmosphere." Note that temperature decreases at a uniform rate with increasing altitude. For comparison, temperature typically drops about 2.5 degrees F per 1000 feet.

Figure 1B shows the typical atmosphere when a temperature inversion is present. If the bend is sharp enough, a duct will form. This condition typically occurs over warm bodies of water and is caused by temperature inversion near the surface of the water. Signals present below the duct will be slightly attenuated. However, if you're above the duct, as you might be during hilltop or mountaintop operation, your signal will undergo severe attenuation towards the duct and you may find communications to be impossible — although others *in* the duct may be experiencing good conditions.

Figure 1C shows a typical profile for an elevated duct. In this case, elevated or mountaintop stations at the proper elevation will experience little or no signal attenuation. Those above or below the duct will experience severe attenuation. This profile is very typical of the ducting that occurs between California and Hawaii at the Hawaiian end. It was carefully measured in the late 1950s by U.S. Navy planes flying at various altitudes near the Hawaiian Islands while monitoring beacons from the California coast.

In the mid-1970s several Amateurs observed this characteristic by driving up and down the side of Mauna Loa on the island of Hawaii. This is one of the reasons that propagation beacons are now located just over 8000 feet above sea level on the side of Mauna Loa. An elevated duct could drop over the path and actually touch the surface as shown on **fig. 1D**. This is probably the typical situation at the California end of the trans-Pacific tropo records.

During the ARRL VHF contest in September, 1984, there was strong evidence to suggest that an elevated duct was the cause of the long distance covered. Jim Stewart, WA4MVI, decided to test the theory by flying a private aircraft at various altitudes on the southern end of the openings (in western South Carolina, in grid square EM-84). He monitored the 432.1-MHz signal of W2SZ/1, located on Mount Greylock in western Massachusetts at about 3500 feet above sea level.

The elevated temperature data he took, shown in **fig. 2**, is unmistakably that of elevated ducting. Signals were strongest between 6000 and 8500 feet above sea level. Signals were very weak above 9500 feet and barely readable above 10,000 feet. Signals decreased slowly below 4000 feet. Ironically, the wind speed in the duct was 40 to 45 knots from the east, but 30 knots from the west above the duct! Hurricane Diana was located just off the northeast coast of Florida at the time.

Unlike a metal waveguide, a duct doesn't have a sharp cutoff frequency. It has a rather slow transition from good trapping (low leakage) to poor trapping (high leakage). The cutoff frequency of a duct is the lowest frequency that will support propagation.

Generally speaking, the duct improves as the frequency is increased until the walls of the duct become too irregular for propagation. Furthermore, the cutoff frequency of a surface duct is lower than that of an elevated duct.[6]

Duct heights of 150 to 1500 feet are most common and seem to be optimum for frequencies of 50 MHz through 1500 MHz[3,4]. Thinner ducts that favor the microwave frequencies are present but are harder to use because of their narrow height. The Europeans have been exploiting these narrower ducts for years by operating on beaches and near large bodies of water during the times of the year when these conditions are most likely to occur. By moving your station and antenna height around, an optimum location can be found — if ducts are present.

scattering

Many types of scatter propagation are especially useful on the microwave frequencies but often overlooked by Amateurs. These include, but are not limited to, aurora, tropospheric scatter, aircraft scatter, rain scatter, light-

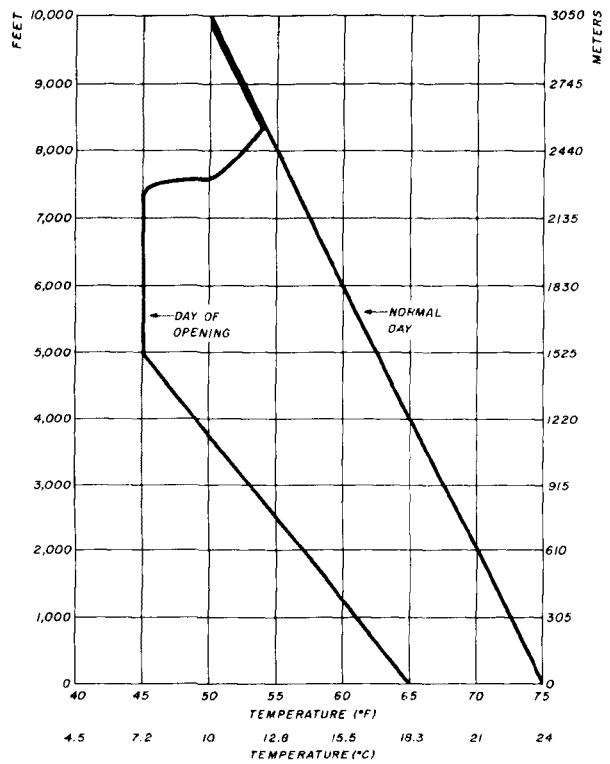


fig. 2. Temperature versus altitude for a typical day as well as data taken at 1400 UTC on September 9, 1984, in the western portion of South Carolina. WA4MVI measured the temperature from an aircraft. Note the breaks at 5000 and 7400 feet (1525 and 2255 meters). See text for further information.

ning scatter, knife-edge diffraction or obstacle gain, and scatter from tall or large objects.

Generally speaking, scatter modes are much more lossy than tropospheric propagation but provide a means to communicate when the normal modes of propagation aren't available. Thus they can provide moderate to long DX more regularly when properly understood and exploited. Because they're more lossy, use better equipment such as higher power and greater antenna gain (more on this later).

Aurora. This mode of scatter propagation has been widely exploited on the frequencies below 450 MHz.[4,5] **Table 4** in *reference 5 lists* recent records, although all of them were exceeded in the February 8, 1986 aurora. Professional studies have shown that auroral scatter is usable up through the 13-cm (2450-MHz) band, but I've never heard of any successful Amateur aurora QSOs above 450 MHz.

One of the problems with auroral propagation is frequency spread. As you go higher in frequency, signals are spread out over a wider bandwidth, so wide bandwidth (2 kHz minimum) receivers are desired. Aurora signals on the UHF bands usually sound as if the background noise level is being keyed to a higher level. Therefore, the received SNR is reduced.

Although this can be offset by using higher gain antennas, doing so also has its drawbacks. The higher the frequency, the narrower the "hot spot." Hence it will be more difficult to locate the optimum reflection point. Add this to the difference in geometry with the station to be worked and you have some fairly tight windows to match.

One of the parameters most overlooked by Amateurs operating auroral scatter is doppler shift. During the auroral opening on February 8, 1986, doppler shifts of up to 4 kHz were observed at 432 MHz. This is clearly out of the tuning range of the RIT on many modern transceivers! Needless to say, there was much QRM and many incomplete QSOs.

To work aurora on the microwave bands, look for a high MUF. This usually happens in the early portion of the opening, especially when radio station WWV is broadcasting a K index of 5 or higher. Other operating tips for detecting aurora are described in detail in references 4 and 5.

Because of the difficulty of properly aiming your antenna, time-sequenced schedules with stations on or within a few degrees of your latitude are recommended. The higher the power, the better — up to the legal limit — with a moderate (20 dBi maximum) antenna gain to maximize the chances of hitting the hot spot. Keep your transmitter frequency fixed and listen up to ±20 kHz! Let me know if you complete a QSO above 450 MHz so we can get you into the record book.

Tropospheric scatter. This mode of propagation, widely used in the commercial and government sectors, is often referred to as *troposcatter*. Huge billboard-type antennas are often used with high power for very high reliability communications over reasonable distances (120 to 620 miles) in the 400 to 5000-MHz frequency range. Even wideband systems such as multiplexed telephone links are used.

The principle of troposcatter is quite straightforward. There are numerous dust particles, clouds, and refractive index changes in the troposphere (1000 to 5000 feet above sea level) that can reflect or scatter microwave signals. If two stations want to communicate, all they have to do is to aim their antennas at a common volume or *scattering* spot as shown in **fig. 3.** Amateur troposcatter communications on frequencies even as high as 3 cm (10 GHz) have been successful, even with moderate (10 watts) power levels!

For best results the beamwidth of the antenna should be moderate (at least a few degrees) because too narrow a beamwidth will miss some of the returns from the scattering volume. Ideally, both stations should use approximately the same beamwidth. Antenna elevation angles should be higher for shorter distances and lower (1 to 2 degrees) for longer distances. For best results there should be a clear horizon because any obstructions will severely attenuate the signals.

Troposcatter can be quite reliable and yield good DX, especially on the microwave bands, but is seldom mentioned by Amateurs. The path loss is roughly equivalent to EME operation for the longer distances just mentioned. New low-noise GaAs FETs and recent developments in higher solid-state power amplifiers make antenna-mounted gear useful in decreasing feedline losses. Further information on troposcatter communications can be found in reference 7.

Aircraft scatter. References 4 and 5 mentioned the tremendous possibilities for QSOs using aircraft scatter out to 500 miles. As the frequency increases, so does the effectiveness of the scattering medium — i.e., the aircraft. Hence there are many possibilities for moderate microwave DX, especially if you live on a path where there are frequent flights, especially at higher altitudes such as in the midwest.

Communicating via aircraft scatter does require timing and patience. Because most aircraft travel at sufficiently high speeds, there will also usually be doppler shift. I have seen shifts approaching 1 kHz at 2304 MHz.

One other caution: aircraft are still quite usable even after they move off the direct path, but *only* if you and the station being worked both follow the aircraft by moving your antenna positions accordingly. I've observed azimuth changes up to 5 degrees at 13 cm; these can be catastrophic if high gain antennas are used. This microwave propagation mode represents an exciting possibililty for year-round communications.

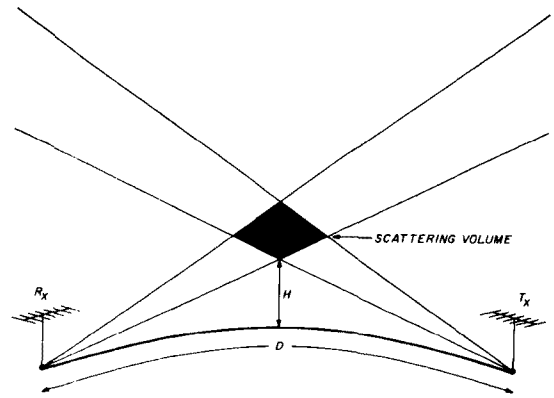


fig. 3. Geometry involved for conducting troposcatter communications. "H" is the height of the scattering medium and "D" is the Great Circle distance between the stations.

Lightning scatter. The possibilities of lightning scatter QSOs were described in reference 4. To my knowledge, this mode has been exploited only through 450 MHz, but should be usable through the microwave frequencies.

As described in reference 4, both stations should aim at the storm center. In this respect, communication will be sort of a cross between aurora and meteor burst because it will occur in bursts that coincide with the lightning itself. Hence communications may be erratic and doppler shift will be quite evident.

Although this mode of propagation has received very little attention, it should yield exciting opportunities for stations located in regions where lightning is quite prevalent, such as in the Gulf states and in the Rocky Mountains.

Rain scatter. This mode of propagation was only recently discovered in the UK.[4] Amateurs aiming their antennas in the direction of a severe rain storm noticed scatter communications over paths where normal communications were not possible.

Best results have been on 3 cm. In some cases one of the two stations involved was aiming straight up as a storm cell passed overhead. This mode should be possible at 5.6 GHz, but tests are inconclusive.

Millimeter propagation via rain scatter will probably not be possible, for reasons to be explained shortly. While this isn't a propagation mode for those using Gunnplexers, it should be quite exciting for well-equipped stations using narrowband communications gear. Stations with obstructed paths — and especially stations not situated at high locations — are particularly favored, especially if they can elevate their antennas.

obstacle gain

We often think that obstacles such as hills or mountain tops are impossible barriers to microwave and millimeter-wave signals, but this isn't true. Experiments have shown that if the peak on a mountain or hill is reasonably sharp, it can reradiate signals in a propagation mode called "knife edge diffraction"[4].

Commercial microwave communications stations have used this mode for many years. The sharper the peak and the closer it is to one end of the path, the more efficient it will be for diffraction. This is an ideal setup for stations located in valleys or those having obstacles that would limit LOS propagation.

Tall buildings, structures, hills, and mountains can also be used to scatter microwave and millimeter-wave signals. If no LOS path is available, but

a tall structure or hill is present, both stations should aim at the structure or hill and scatter their signals that way. 23 cm and up seem to be ideal for this mode.

EME communications

This is another form of scatter communications. The path loss is typically 275 dB at 1296 MHz and proportionally higher as you go up in frequency. The background or sky noise temperature above 900 MHz is very low. A good low-noise parabolic dish, 12 to 15 feet in diameter, with a low-noise GaAs FET preamplifier and 100 watts, is more than sufficient to produce echoes at 1296 MHz and above. Furthermore, if the antenna size is kept constant, the signal strength will actually increase as you go higher in frequency!

The highest reported Amateur QSO via EME is at 2320 MHz. There are no reports yet in the new 33-cm band. The possibilities of EME are endless because there are typically 8 to 11 hours of moon-time every day of the year.

EME communications also yields a platform for testing new antennas, preamplifiers, and transmitters; you're able to listen to your improvements 2.5 seconds later! No more wondering if your gear is functional or if the band is open but no one's there. You *know* that path is open! EME offers a great challenge and is within Amateur capabilities at least through 3 cm!

millimeter-wave propagation

So far, most of the propagation modes mentioned apply mainly to microwaves. Millimeter-wave — or the world above 3 cm — is definitely more challenging than the microwave frequencies. Tropospheric and some of the scatter modes of propagation are available but more difficult to explore fully. With the millimeter-wave gear presently available, most operation depends upon LOS propagation.

But it won't always be easy. Why? Because as you go above 20 GHz, you approach the resonance frequencies of water and oxygen. As a result, there may be unexpected attenuation due to absorption, especially on certain frequencies, *some of which are in or near our Amateur bands.*

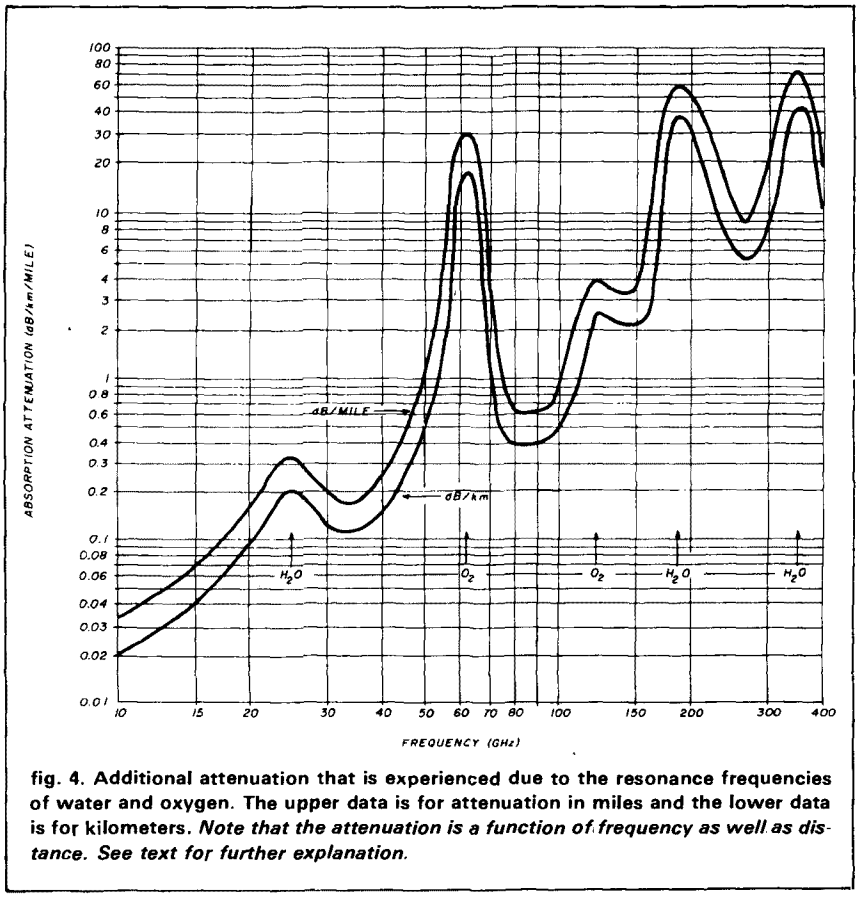


fig. 4. Additional attenuation that is experienced due to the resonance frequencies of water and oxygen. The upper data is for attenuation in miles and the lower data is for kilometers. *Note that the attenuation is a function of frequency as well as distance. See text for further explanation.*

Furthermore, fog and rain have a pronounced effect on millimeter-wave band performance. Two things must be remembered for improving millimeter-wave operation: first, for greatest distances, operate on a day when the humidity is low. This probably means the late fall or even winter — definitely not the time of year during which you'd normally expect to set DX records. Second, as you go higher in altitude, absorption attenuation and humidity usually drop, so mountaintopping is highly recommended.

Figure 4 is useful for determining the additional path loss due to water and oxygen absorption. Note that the additional path loss is not only a function of frequency but also of path length. Therefore, in order to determine the nominal loss on a millimeter-wave path, you have to first determine the normal LOS path loss using **eqns 1** and **2** or **fig. 1** in reference 1 and then add this to the loss attributable to oxygen and water absorption.

For example, let's assume that you're calculating the LOS path loss for a 20-mile link on 24 GHz. Using reference 1, the nominal path loss is approximately 151 dB. **Figure 4** shows that the additional loss due to absorption is approximately 0.32 dB per mile because of the proximity of water resonance (frequency). So the additional attenuation is 6.4 dB for a total LOS path loss of approximately 157.4 dB. As a rule of thumb, the absorption attenuation will be about one-tenth the sea level value shown in **fig. 4** at 30,000 feet elevation. Anyone for trying Mount Everest?

The absorption losses at 48 GHz are

about the same as at 24 GHz, although the LOS path loss is 6 dB higher. However, on our new 4-mm (76 GHz) band the absorption losses would be about 5 times higher! This is true because the latter band is close to the resonance frequency of oxygen.

The 75 GHz band, then, is a real challenge to Amateurs. Late-breaking news is that the first known Amateur QSO on 75 GHz took place on December 30, 1985, between HB9AGE and HB9MIN at a distance of 0.3 miles. Congratulations!

passive reflectors

Special reflectors can be used to great advantage on the upper microwave and millimeter-wave bands. They can scatter signals in a manner similar to that of obstacle gain just discussed.

However, while mountains can't be easily moved, passive reflectors can be optimized because the direction can usually be controlled. One type of reflector, often called a "billboard," is commonly used in the commercial community in situations in which mountains or hills must be spanned but the equipment can't be mounted high enough. As with knife-edge diffraction, the optimum solution is to mount the reflector closer to one station rather than at or near the middle of the path.

A special variation of the passive reflector popular among European Amateurs is the "flyswatter" or "periscope" antenna.[8,9,10] Often used on 3 cm and above, its active equipment is mounted near the ground. Its power is radiated from a horn or small parabolic dish antenna aimed directly at a large reflecting surface placed at a 45-degree angle between 50 and 100 feet above this radiator. The reflective surface reradiates most of the RF inthe desired direction. This antenna configuration is particularly recommended for stations that would normally require long runs of transmission line.

repeaters

Although this particular column addresses weak-signal work, I would be remiss if I didn't at least mention active repeaters. Let's face it: FM repeaters are everywhere. For technical reasons, it's often desirable to separate the receiver and transmitter by a good distance, so there's the need for a link between the two sites.

Low-power links are usually all that's required to connect the repeater receiver and transmitter. Microwave links, even those as simple as Gunnplexers, are ideal for this type of communications. So when you access a 2-meter FM repeater, you may be using the microwave bands without even knowing it!

major obstacles to communications

So far I've been stressing the positive possibilities of microwave and millimeter-wave propagation. But to be honest, communication on the microwave and millimeter-wave frequencies has its shortcomings — but these are usually a challenge to the enterprising Amateur.

Ice, rain, and fog surely present problems, especially on the millimeter-wave bands as just discussed. Try to work around them by choosing clear, dry weather or by operating from a high location when attempting new DX records.

Foliage attenuation — even that caused by evergreens — can totally disrupt a microwave path. In fact, the additional attenuation from deciduous and evergreen foliage can even be noticed at as low as 2 meters! On 23 cm and above, pine needles act like reradiating dipoles that significantly increase the attenuation on a particular path. Operation in the winter is no real cure-all, either, since most of the good tropo types of propagation aren't prevalent at this time of year. Recently there's been a lot of interest in foliage attenuation, especially in mobile communications and on the microwave bands. In fact, some typical data is just beginning to be correlated. It's been suggested that between 200 MHz and 95 GHz, attenuation can be calculated — if the depth of a grove of trees to be penetrated is less than 1200' feet — using the following equation:

$$L = 0.187 \bullet f^{0.284} \bullet D^{0.588} \quad (1)$$

where L is loss in dB, f is frequency in MHz, and D is depth of the grove expressed in meters. The CCIR Study Group recommends a more simplified equation:

$$L = 0.2 \bullet f^{0.3} \bullet D^{0.6} \quad (2)$$

For example, using a scientific calculator, the attenuation through a 10-meter grove of trees at 1 GHz would be about 5.15 dB using **eqn 1** and 6.32 dB using **eqn 2**. Although there are differences, this may be the first time that Amateurs may be able to quantitatively determine the destructive effects of foliage attenuation.

Before leaving the subject of obstacles to propagation, it may be appropriate to again review the subject of antenna height. I've pointed out many times that the maximum height for Amateur antennas on VHF and above should be 2 to 3 wavelengths, or just above any local obstructions, whichever is higher.

This factor is particularly true as frequency increases. A taller tower requires a longer transmission line that can often increase attenuation at a faster rate than the increased gain attributable to greater antenna height. The overall loss of transmitter power is important, but the loss in the received signal is often more important, since the noise levels are usually so low above 1 GHz.

summary

This two-part series was prepared to give an overall view of microwave and millimeter-wave radio propagation and how to best exploit DXing on those bands. Many of the typical propagation modes — and some of the lesser known ones — were discussed. Light waves were not discussed because they're a subject in themselves.

Because the majority of radio propagation conditions that facilitate DX operation are called "anomalous" or "disruptive" by commercial and government users, most research has been done from a negative point of

view. It is up to Amateurs to see the positive side of this data.

Many of the references used in preparation of this column weren't detailed because they are cited in the references listed, particularly in references 4 and 5.

acknowledgements

I'd like to thank Jim Stewart, WA4MVI, for sharing his meteorolgical data on the now-famous east-coast tropo duct of September, 1984.

See you soon on the microwave and millimeter-wave bands!

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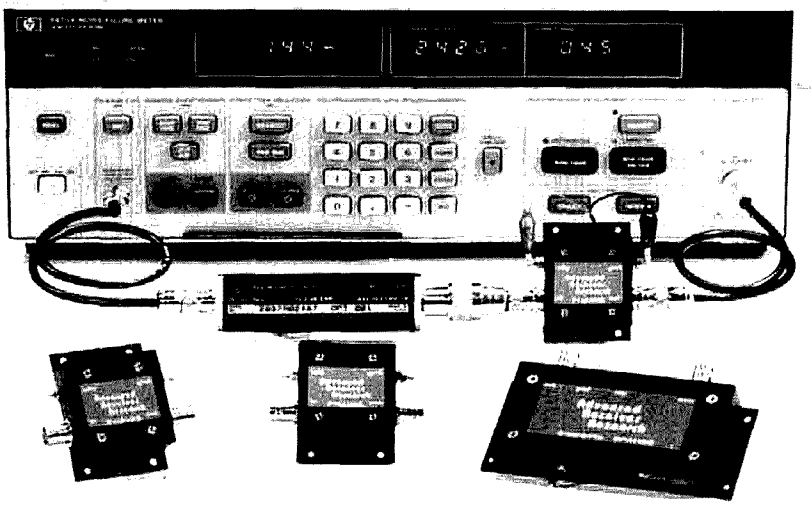
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important VHF/UHF events:

August 2-3	*ARRL UHF Contest*
August 12	*Predicted peak of the Perseids meteor shower at 0700 UTC.*
August 16	*EME perigee*
August 28-31	*"Microwave Update 1986 Conference," Estes Park, Colorado (contact W0PW).*
September 6-7	*International Region 1 VHF Contest.*
September 12	*EME perigee*
September 13-14	*ARRL VHF QSO Party.*
September 26-27	*First ARRL 10 GHz Cumulative Contest.*

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P144VDA	144-148	<1.0	15	0	DGFET	$37.95
P144VDG	144-148	<0.5	24	+12	GaAsFET	$79.95
P220VD	220-225	<1.8	15	0	DGFET	$29.95
P220VDA	220-225	<1.2	15	0	DGFET	$37.95
P220VDG	220-225	<0.5	20	+12	GaAsFET	$79.95
P432VD	420-450	<1.8	15	−20	Bipolar	$32.95
P432VDA	*420-450*	*<1.1*	*17*	*−20*	*Bipolar*	$49.95
P432VDG	420-450	<0.5	16	+12	GaAsFET	$79.95
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SP28VD	28-30	<1.2	15	0	DGFET	$59.95
SP50VD	50-54	<1.4	15	0	DGFET	$59.95
SP50VDG	50-54	<0.55	24	+12	GaAsFET	$109.95
SP144VD	144-148	<1.6	15	0	DGFET	$59.95
SP144VDA	144-148	<1.1	15	0	DGFET	$67.95
SP144VDG	144-148	<0.55	24	+12	GaAsFET	$109.95
SP220VD	220-225	<1.9	15	0	DGFET	$59.95
SP220VDA	220-225	<1.3	15	0	DGFET	$67.95
SP220VDG	220-225	<0.55	20	+12	GaAsFET	$109.95
SP432VD	420-450	<1.9	15	−20	Bipolar	$62.95
SP432VDA	420-450	<1.2	17	−20	Bipolar	$79.95
SP432VDG	420-450	<0.55	16	+12	GaAsFET	$109.95

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how to plot great circles
on your favorite map

Bearings at a glance for the Amateur DXer

Many active hams can recall the tremendous activity of the FCC in the years just before and during World War II. At the time, the authors were engineers engaged in developing a system for intercepting unauthorized, possibly subversive, radio transmissions and attempting to determine their source. These activities were part of the wartime mission of the FCC's Radio Intelligence Division (RID), under the direction of George E. Sterling, W1AE, and the late Charles A. Ellert, W3LO. Even before the United States entered the war, RID had developed a nationwide system of monitoring and direction finding for these purposes.

Shortly after the bombing of Pearl Harbor, the military requested that the FCC assume responsibility for radio security of the Territory of Hawaii. It also asked that a direction-finding system be established to provide navigational information for lost military aircraft flying over the Pacific Ocean, generally from the west coast to Hawaii. Six Adcock DF stations were located at strategic sites on Oahu, Kauai, Maui, Molokai, and the "big island," Hawaii. These stations maintained two-way radio communication with the Radio Security Center in Honolulu, where bearings on lost aircraft transmissions were plotted and evaluated, with the results then transmitted to Hickam Air Force Base.

At the Radio Security Center, bearings were plotted on a Gnomonic Projection map of the central Pacific area. Individual "compass roses" for each Adcock station were computed, using the meridian intercept system developed by Charles Ellert. It is a tribute to Sterling and Ellert that the Hawaiian system, with its DF bearings, safely brought in 272 lost bombers, hospital aircraft, and passenger planes during its two years of operation.

Computation of the compass roses for the DF stations was a laborious process. The hand calculator and computer had not yet been developed, and slide-rule calculations weren't sufficiently accurate, so logarithm tables had to be used. This article describes W3LO's system for computing great circle tracks using the meridian intercept method. All computation can be done with either a calculator or a computer; both methods are fully described.

making the map

Referring to a computer readout of bearings to each country can be slow and cumbersome when you're eager to snag a new one. It's much quicker and easier to refer to a map that shows great circle bearings from your location to all others throughout the world.

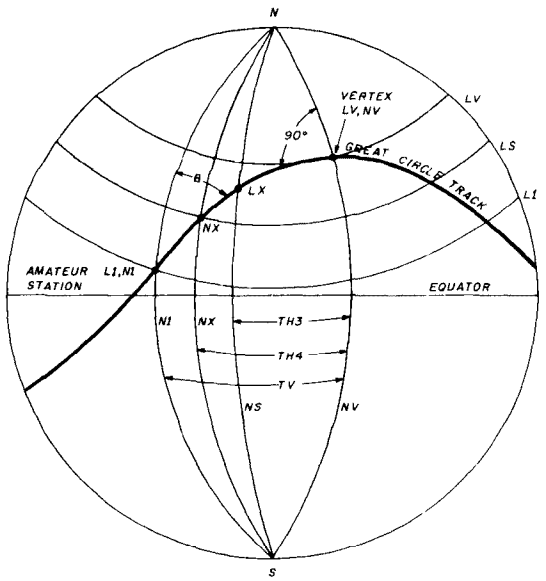
spherical geometry and maps

The earth is an oblate spheroid — a slightly flattened sphere. Maps are, of course, intended to represent all or a portion of the earth's surface; but the greater the area of the earth included, the greater the difficulty in achieving a realistic portrayal of the earth. (Try cutting a hollow rubber ball in half and flattening it . . . you'll get a good idea of the problem. Considerable distortion and deformation obviously result).

Latitude and longitude lines, the imaginary reference marks used to locate points on a map, correspond to equivalent points on the earth's surface. It is these lines that define a map's projection. But most maps are not truly projections in the geometric sense.

Several kinds of maps are commonly found in the radio room. Cartographers have developed special-purpose maps such as the Gnomonic, Lambert Conformal Conic, Polyconic, Albers Equal-Area, Azimuthal, and the familiar projection developed in 1538 by Gerhard Kramer, better known by his Latin sur-

By A. Prose Walker, W4BW, 1087 Tung Hill Drive, Tallahassee, Florida 32301 and **Jack W. Herbstreit, WØDW,** 4797 Briar Ridge Trail, Boulder, Colorado 80301



B	Bearing angle from Amateur station measured clockwise from True North
L1,N1	Latitude and longitude of Amateur station
LV,NV	Latitude and longitude of the Vertex
LS,NS	Selected latitude and longitude lines on which bearing intercepts are computed
LX	Latitude of Bearing B intercept on NS
NX	Longitude of Bearing B intercept on LS
TV	Longitude difference between the Vertex and Amateur station
TH3	Longitude difference between NV and NS
TH4	Longitude difference between NV and NX

fig. 1. Geometric Portrayal of pertinent latitude, longitude lines and great circle track intercepts for computation of Great Circle Tracks using the Meridian Intercept Method.

name, Mercator. An equal-area projection, the Mercator projection is simple to construct. It uses straight lines at right angles to each other for latitude and longitude. Its scale increases with latitude.

The Gnomonic Projection is the only map on which great circle tracks appear as straight lines. The more common azimuthal maps used by Amateurs are special cases of the Gnomonic. For radio bearings (i.e., azimuths) over considerable distances, this map has no equal. Even so, there are distortions (remember the rubber ball?), and individual "compass roses" must be computed for each station's location, except at the point of tangency. The Gnomonic is a perspective projection of the earth on a tangent plane, with the point from which the projecting lines are drawn being situated at the center of the earth.

Regardless of the map used, the path of a great circle track around the world will inevitably intercept latitude and longitude lines as it passes from the origin (station location) to the antipode. The object of this "meridian intercept" system is to determine where the great circle paths intercept selected parallels and meridians of latitude and longitude so that they can be plotted on a map.

the meridian intercept system

The diagram shown in **fig. 1** is the basis for the computations of intercepts using a hand calculator for solving the equations. The corresponding computer program is shown in **fig. 2**. Studying **fig. 1** and referring to a globe will help in visualizing the process more easily.

Although they may appear to be slightly different, the equations for the calculator* are identical to those used in the computer program. Although somewhat tedious, they do provide accurate results. The station location need be specified only to the nearest degree of latitude and longitude; this helps simplify calculation. Map distortion, plotting accuracy, and horizontal beamwidths of Amateur directional antennas make extreme accuracy unnecessary.

intercept calculations using a calculator

In **fig. 1**, coordinates of the Amateur station are L1, N1. The bearing, B, is represented by a heavy black line. Its great circle track reaches a maximum latitude at a point called the vertex, with coordinates LV, NV. For each great circle track there are two vertices, one east and one west of the Amateur station. If intercepts are computed for bearings from 0 to 180 degrees, reciprocals of those intercepts may be used for bearings 180-360 degrees. It's necessary to compute the latitude and longitudes of the vertex so that values for TV, TH3, TH4, LX, and NX may be obtained. Take time to understand the definitions in **fig. 1** — they're important.

Finding LV: latitude of the vertex
Example: let B = 60 degrees from W4BW
L1 = 30.4N (latitude of W4BW)
N1 = 84.2W (longitude of W4BW)

$$COS(LV) = SIN((PI/2) - L1) \cdot SIN(B) \quad \textbf{(eqn 1)}$$

solving: $COS(LV) = 0.86251 \cdot 0.86603$
$COS(LV) = 0.74696,\ LV = 41.7$ degrees

Finding NV: longitude of the vertex (using definitions value of TV and N1)

$$COT(TV) = COS((PI/2)-L1) \cdot TAN(B) \quad \textbf{(eqn 2)}$$

using value from above example:

*The original "Ellert equations" are given below. In all cases they give identical results to those used herein. When using a hand calculator, these original equations may be more convenient. With these equations (or the others) it is nesessary to carefully ensure that the answers obtained have the correct signs and fall in the correct quadrants on the map of your choice. Roughly plotting your intercepts will verify the correct quadrant. It is useful to have your map available when making the calculations.

$$COS(LV) = SIN(B) \cdot COS(L1) \quad \textbf{(eqn 1)}$$
$$COS(TV) = SIN(L1) \cdot TAN(B) \quad \textbf{(eqn 2)}$$
$$TAN(LX) = COS(TH3) \cdot TAN(LV) \quad \textbf{(eqn 3)}$$
$$COS(TH4) = TAN(LS) \cdot COT(LV) \quad \textbf{(eqn 4)}$$

$$COT(TV) = 0.50603 \cdot 1.73205$$

COT (TV) = 0.87647,

TV = 48.8 degrees

TV, by definition, is the longitude difference between the station and the nearest vertex: consequently

$NV = N1 \quad TV$

or $NV = 84.2 - 48.8 = 35.4W$ *longitude of the vertex (NV).*

Before proceeding, note the definition of TH3. Select suitable (NS) meridians that a bearing of 60 degrees (B) will intercept. A different value of LX will result for each selected meridian. In the following example, values of NS used are 30W and 60E.

$$TAN(LX) = COS(TH3) \cdot TAN(LV) \qquad \textbf{(eqn 3)}$$

Example:

NS	= 30W	60E
TH3	= 5.4	95.4
COS(TH3)	= 0.99556	−0.09411
TAN(LV)	= 0.89097	0.89097 (LV from Eqn. 1)
TAN(LX)	= 0.88701	−0.08385 (reverse latitude)
LX	= 41.5N	4.8S

In many cases, use of **eqn 4** may not be necessary. In other words, the latitude intercepts on longitude meridians will generally locate the great circle points with sufficient clarity to enable them to be joined with a smooth curve. In some circumstances, such as when the great circle track approaches the antipode, accurate location of the points may require solving **eqn 5** to obtain longitude intercepts NX, on several selected latitudes NS. Note again the definitions from **fig. 1** before setting up your table for **eqn 4**. In the following example, azimuths B of 10, 20, 30, and 40 degrees are used to compute the longitude intercepts NX on a latitude LS of 60 degrees.

$$COS(TH4) = COT((PI/2) - LS) \cdot TAN((PI/2) - LV) \qquad \textbf{(eqn 4)}$$

B	10	20	30	40
LS	60N	60N	60N	60N
LV	81.4N	72.8N	64.5N	56.3N
TAN((PI/2) - LV)	0.15124	0.30955	0.47698	0.66692
COT((PI/2)-LS)	1.73205	1.73205	1.73205	1.73205
COS(TH4)	0.26196	0.53616	0.82615	(1.15519)*
TH4	74.8	57.6	34.1	. . . . .
NV	−0.7	4.6	10.5	17.2
NX	74.1	62.2	44.6	no intercept

max value of cosine is 1 or −1

Thus we have computed longitude intercepts on a 60-degree north latitude for bearings of 10, 20, and 30 degrees. Reference to your map will verify that there is no intercept for a bearing of 40 degrees on

```
5 LPRINT "A. Prose Walker, W4BW and Jack W. Herbstreit, WØDW":LPRINT :LPRINT
10 LPRINT "                    VERTEX"; :LPRINT TAB(66) DATE$
20 LPRINT "PROGRAM TO COMPUTE CROSSINGS OF GREAT CIRCLE PATHS ON MERIDIANS"
30 LPRINT "OF LATITUDES AND LONGITUDES FROM A POINT ON THE EARTHS SURFACE"
40 LPRINT "AT SELECTED BEARINGS, B,FROM THE POINT OF ORIGIN."
50 '
60 PI=3.141593
70 DEF FNR(X1)=X1*PI/180
80 DEF FND(X2)=X2*180/PI
90 INPUT "LATITUDE OF POINT OF ORIGIN IN DEGREES NORTH OF THE EQUATOR ";L1
100 INPUT "LONGITUDE OF POINT OF ORIGIN IN DEGREES WEST OF GRNWCH. ";N1
110 LPRINT "GIVEN THE LATITUDE";L1;"DEGREES AND LONGITUDE ";N1;"DEGREES OF THE"
120 L1=FNR(L1)
130 N1=FNR(N1)
140 LPRINT "POINT OF ORIGIN (DEGREES WEST OF GREENWICH), NORTH LATITUDES ARE"
150 LPRINT "GIVEN AS POSITIVE AND SOUTH LATITUDES ARE NEGATIVE, WEST"
160 LPRINT "LONGITUDES ARE GIVEN AS POSITIVE AND EAST LONGITUDES ARE NEGATIVE."
170 LPRINT "B IS SELECTED BEARING FROM POINT ON EARTHS SURFACE IN DEGREES "
180 LPRINT: LPRINT
190 LPRINT "LATITUDE OF VERTEX (HIGHEST LATITUDE THAT BEARING CROSSES) = LV"
200 LPRINT "LONGITUDE OF VERTEX EAST OF ORIGIN = NVE"
210 LPRINT "LONGITUDE OF VERTEX WEST OF ORIGIN = NVW"
220 LPRINT "LONGITUDE DIFFERENCE BETWEEN ORIGIN AND NEAREST VERTEX = TV"
230 LPRINT
240 'LS=FNR(LS) SELECTED LATITUDES ON WHICH BEARING CROSSES ARE COMPUTED
250 'NS=FNR(NS) SELECTED MERIDIANS ON WHICH BEARING CROSSES ARE COMPUTED
260 'LX=FNR(LX) LATITUDES OF CROSSINGS ON NS
270 'NX=FNR(NX) LONGITUDES OF CROSSINGS ON LX
280 'TV=(N1-NV) LONGITUDE DIFFERENCE BETWEEN NEAREST VERTEX AND POINT OF ORIGI
290 DIM A(19,39)
300 FOR I = 0 TO 18
310     C = I * 10
320 A(I,1)=C
330 C = FNR(C)
340 CLV=SIN((PI/2)-L1)*SIN(C)
350 LLV=ATN(CLV/SQR(1-CLV*CLV)) 'LATITUDE OF VERTEX
360 LV=(PI/2)-CLV
370 A(I,2)=LV
380 IF TAN(C)=0 THEN TV=PI/2:GOTO 410
390 TTV=1/((COS((PI/2)-L1))*TAN(C))
400 TV=ATN(TTV)
410 A(I,4)=TV
420 NVE=N1-TV 'LONGITUDE OF VERTEX EAST OF ORIGIN
430 A(I,3)=NVE
440 NVW=N1+TV 'LONGITUDE OF VERTEX WEST OF ORIGIN
450 A(I,4)=NVW
460 A(I,5)=TV
470 FOR J = -6 TO 5
480     NS = J * 30
490 NS = FNR(NS) 'SELECTED LONGITUDES
500 TH3E=NVE-NS 'LONGITUDE DIFFERENCE BETWEEN NVE (EASTERN VERTEX) AND NS
510 TH3W=NVW-NS 'LONGITUDE DIFFERENCE BETWEEN NVW (WESTERN VERTEX) AND NS
520 LXE=COS(TH3E)*TAN(LV)
530 LXW=COS(TH3W)*TAN(LV)
540 LXE=ATN(LXE) 'LATITUDE OF CROSSING EAST OF VERTEX ON SELECTED LONGITUDE, NS
550 LXW=ATN(LXW) 'LATITUDE OF CROSSING WEST OF VERTEX ON SELECTED LONGITUDE, NS
560 A(I,J+12)=LXE
570 NEXT J
580 '
590 FOR J = -5 TO 5
600         LS = J * 15
610 LS = FNR(LS)
620 TH4=(ATN((PI/2)-LV)/TAN((PI/2)-LS)
630 IF ABS(TH4)>1 THEN GOTO 670
640 NXEW=0
650 NXWE=0
660 GOTO 790
670 TH4=(-ATN(TH4/SQR(1-TH4*TH4)))+PI/2
680  TH4=LONGITUDE DIFFERENCE BETWEEN NV AND NX
690 NXEE=NVE-TH4 'LONGITUDE OF CROSSING EAST OF EASTERN VERTEX
700 NXEW=NVE+TH4 'LONGITUDE OF CROSSING WEST OF EASTERN VERTEX
710 NXWE=NVW-TH4 'LONGITUDE OF CROSSING EAST OF WESTERN VERTEX
720 NXWW=NVW+TH4 'LONGITUDE OF CROSSING WEST OF WESTERN VERTEX
730 IF NXEW<PI THEN NXEW = NXEW - 2*PI: GOTO 750
740 IF NXEW>-PI THEN NXEW = NXEW + 2*PI
750 A(I,J+23)=NXEW
760 IF NXWE<PI THEN NXWE = NXWE - 2*PI: GOTO 790
770 IF NXWE>-PI THEN NXWE = NXWE + 2*PI
780 A(I,J+34)=NXWE
790 NEXT J
800 NEXT I
810 LPRINT TAB(8) "TABLE I": LPRINT
820 LPRINT TAB(2) "B"; TAB(7)"LV"; TAB(12) "NVE"; TAB(18) "NVW"; TAB(25)"TV"
830 FOR I= 0 TO 18
840 LPRINT USING "###";A(I,1);
850 LPRINT USING "####.#";FND(A(I,2));
860 LPRINT USING "####.#";FND(A(I,3));
870 LPRINT USING "####.#";FND(A(I,4));
880 LPRINT USING "####.#";FND(A(I,5))
890 NEXT I
895 LPRINT CHR$(12)
896 LPRINT "A. Prose Walker, W4BW and Jack W. Herbstreit, WØDW":LPRINT :LPRINT
900 LPRINT TAB(30) "TABLE II":LPRINT
910 LPRINT"LATITUDE  OF CRUSSING FOR DESIGNATED LONGITUDE BETWEEN 180 AND 150W"
920 LPRINT"B/LO";"  180";"  E150";"  E120";"  E90  ";"  E60";"   E30";"     0";
930 LPRINT "   30W";"   60W";"   90W";"  120W";"  150W"
940 FOR I = 0 TO 18
950 LPRINT USING "###";A(I,1);
960 FOR J = 1 TO 11
970 LPRINT USING "####.#";FND(A(I,J+5));
980 NEXT J
990 LPRINT USING "####.#";FND(A(I,17))
1000 NEXT I
1010 LPRINT :LPRINT  TAB(30) "TABLE III": LPRINT
1020 LPRINT "LONGITUDE AT CROSSING FOR DESIGNATED LATITUDES BETWEEN -75 AND 75"
1030 LPRINT"B/LA";"  S75";"  S60";"   S45";"   S30";"   S15";"     0";"   15N";
1040 LPRINT "   30N";"   45N";"   60N";"   75N"
1050 FOR I = 0 TO 18
1060 LPRINT USING "###";A(I,1);
1070 FOR J = 1 TO 10
1080 LPRINT USING "####.#";FND(A(I,J+17));
1090 NEXT J
1100 LPRINT USING "####.#";FND(A(I,28))
1110 NEXT I
1115 LPRINT CHR$(12)
1116 LPRINT "A. Prose Walker, W4BW and Jack W. Herbstreit, WØDW":LPRINT :LPRINT
1120 LPRINT :LPRINT  TAB(30) "TABLE IV":LPRINT
1130 LPRINT "LONGITUDE AT CROSSING FOR DESIGNATED LATITUDES BETWEEN -75 AND 75"
1140 LPRINT"B/LA";"  S75";"  S60";"   S45";"   S30";"   S15";"     0";"   15N";
1150 LPRINT "   30N";"   45N";"   60N";"   75N"
1160 FOR I = 0 TO 18
1170 LPRINT USING "###";A(I,1);
1180 FOR J = 1 TO 10
1190 LPRINT USING "####.#";FND(A(I,J+28));
1200 NEXT J
1210 LPRINT USING "####.#";FND(A(I,39))
1220 NEXT I
1225 LPRINT CHR$(12)
1230 END
```

fig. 2. Program listing to determine great circle paths from your location. (Send SASE for large-print copy of program.)

the 60th parallel because the great circle track passes south of that latitude.

The principles and notations for determining intercepts with a calculator are the same as will be later explained for the computer program (see **fig. 2**). LS and NS values are selected as computations progress, taking into account the easily identified *parallels* and *meridians* as shown on a Mercator map ("Standard Time Zone Chart of the World," HO 5192, published by the United States Naval Oceanographic Office). Because of the time zones, some of them were not even values. In the computer program, bearings (B) every 10 degrees are used; selected longitude lines (NS) are every 30 degrees, and selected latitudes (LS,) every 15 degrees from 75S to 75N. The photo of the W4BW map, **fig. 3**, shows the great circle paths throughout the world from Tallahassee, Florida.

table 1. Latitudes and longitudes of the vertices for W4BW.

B	LV	NVE	NVW	TV
0	90.0	-5.8	174.2	90.0
10	81.4	-0.7	169.1	84.9
20	72.8	4.6	163.8	79.6
30	64.5	10.5	157.9	73.7
40	56.3	17.2	151.2	67.0
50	48.6	25.3	143.1	58.9
60	41.7	35.4	133.0	48.8
70	35.9	48.5	119.9	35.7
80	31.9	65.0	103.4	19.2
90	30.4	84.2	84.2	-0.0
100	31.9	103.4	65.0	-19.2
110	35.9	119.9	48.5	-35.7
120	41.7	133.0	35.4	-48.8
130	48.6	143.1	25.3	-58.9
140	56.3	151.2	17.2	-67.0
150	64.5	157.9	10.5	-73.7
160	72.8	163.8	4.6	-79.6
170	81.4	169.1	-0.7	-84.9
180	90.0	-5.8	174.2	90.0

Program to compute crossings of great circle paths on meridians of latitudes and longitudes from a point on the earth's surface at selected bearings, B, from the point of origin. Given the latitude 30.4 degrees and longitude 84.2 degrees of the point of origin (degrees west of Greenwich). North latitudes are given as positive and south latitudes are negative. West longitudes are given as positive and east longitudes are negative. B is selected bearing from point on earths surface in degrees.
Latitude of vertex (highest latitude that bearing crosses) = LV
Longitude of vertex east of origin = NVE
Longitude of vertex west of origin = NVW
Longitude difference between origin and nearest vertex = TV

program vertex

The computer program uses simple right-angle spherical trigonometric relationships and computes the crossings of great circle tracks from any origin (for example, your location at latitude L1 and longitude N1) on parallels of latitude and meridians of longitude on your map. As with the calculator, select suitable latitude and longitude lines around the world wherever you want the bearing to be plotted (on any type of map of the world, such as a

table 2. Latitudes of crossings for designated longitudes between 80 and 150 w.

B/LO	180	E150	E120	E90	E60	E30	0	30W	60W	90W	120W	150W
0	90.0	90.0	90.0	-90.0	-90.0	-90.0	-90.0	-90.0	-90.0	90.0	90.0	90.0
10	-81.4	-80.0	-72.8	4.6	73.5	80.1	81.4	80.0	72.8	-4.6	-73.5	-80.1
20	-72.8	-71.1	-61.5	-14.7	54.2	69.4	72.8	71.1	61.5	14.7	-54.2	-69.4
30	-64.1	-63.1	-53.6	-20.8	34.9	57.9	64.1	63.1	53.6	20.8	-34.9	-57.9
40	-55.1	-55.7	-47.8	-23.9	18.4	45.6	55.1	55.7	47.8	23.9	-18.4	-45.6
50	-45.8	-48.5	-43.0	-25.9	5.3	32.9	45.8	48.5	43.0	25.9	-5.3	-32.9
60	-36.0	-41.5	-39.0	-27.3	-4.8	20.3	36.0	41.5	39.0	27.3	4.8	-20.3
70	-25.6	-34.4	-35.3	-28.4	-12.9	8.2	25.6	34.4	35.3	28.4	12.9	-8.2
80	-14.7	-27.0	-31.8	-29.4	-19.6	-3.1	14.7	27.0	31.8	29.4	19.6	3.1
90	-3.4	-18.9	-28.2	-30.3	-25.4	-13.5	3.4	18.9	28.2	30.3	25.4	13.5
100	8.2	-10.1	-24.3	-31.1	-30.8	-23.1	-8.2	10.1	24.3	31.1	30.8	23.1
110	19.8	-0.1	-19.9	-32.1	-35.9	-32.0	-19.8	0.1	19.9	32.1	35.9	32.0
120	31.2	11.3	-14.6	-33.1	-40.9	-40.4	-31.2	-11.3	14.6	33.1	40.9	40.4
130	42.3	24.0	-7.8	-34.3	-46.3	-48.4	-42.3	-24.0	7.8	34.3	46.3	48.4
140	52.8	37.9	1.8	-35.9	-52.1	-56.3	-52.8	-37.9	-1.8	35.9	52.1	56.3
150	62.7	52.1	16.1	-38.2	-58.8	-64.2	-62.7	-52.1	-16.1	38.2	58.8	64.2
160	72.2	65.9	37.6	-42.2	-66.9	-72.4	-72.2	-65.9	-37.6	42.2	66.9	72.4
170	81.2	78.7	65.2	-51.3	-77.0	-80.9	-81.2	-78.7	-65.2	51.3	77.0	80.9
180	90.0	90.0	90.0	-90.0	-90.0	-90.0	-90.0	-90.0	-90.0	90.0	90.0	90.0

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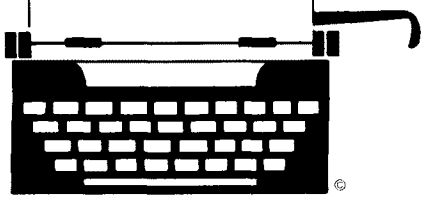
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table 3. Longitudes of great circle track crossings at specific latitudes, using longitude of the eastern vertex.

B/LA	S75	S60	S45	S30	S15	0	15N	30N	45N	60N	75N
0	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2
10	123.7	104.5	98.0	94.3	91.6	89.3	87.0	84.3	80.6	74.1	54.9
20	0.0	127.0	112.6	104.9	99.4	94.6	89.9	84.4	76.7	62.3	0.0
30	0.0	156.4	129.0	116.5	107.8	100.5	93.1	84.5	71.9	44.6	0.0
40	0.0	0.0	149.0	129.8	117.5	107.2	96.9	84.6	65.4	0.0	0.0
50	0.0	0.0	177.0	145.8	128.9	115.3	101.7	84.7	53.6	0.0	0.0
60	0.0	0.0	0.0	165.9	143.0	125.4	107.9	85.0	0.0	0.0	0.0
70	0.0	0.0	0.0	−168.5	160.2	138.5	116.7	85.5	0.0	0.0	0.0
80	0.0	0.0	0.0	−136.7	−179.5	155.0	129.4	86.7	0.0	0.0	0.0
90	0.0	0.0	0.0	−106.0	−158.6	174.2	147.0	94.4	0.0	0.0	0.0
100	0.0	0.0	0.0	−98.3	−141.0	−166.6	167.9	125.1	0.0	0.0	0.0
110	0.0	0.0	0.0	−97.1	−128.3	−150.1	−171.8	156.9	0.0	0.0	0.0
120	0.0	0.0	0.0	−96.6	−119.5	−137.0	−154.6	−177.5	0.0	0.0	0.0
130	0.0	0.0	−65.2	−96.3	−113.3	−126.9	−140.5	−157.4	171.4	0.0	0.0
140	0.0	0.0	−77.0	−96.2	−108.5	−118.8	−129.1	−141.4	−160.6	0.0	0.0
150	0.0	−56.2	−83.5	−96.1	−104.7	−112.1	−119.4	−128.1	−140.6	−168.0	0.0
160	0.0	−73.9	−88.3	−96.0	−101.5	−106.2	−111.0	−116.5	−124.2	−138.6	0.0
170	−66.5	−85.7	−92.2	−95.9	−98.6	−100.9	−103.2	−105.9	−109.6	−116.1	−135.0
180	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2

table 4. Longitudes of great circle track crossings at specific latitudes, using longitudes of the western vertex.

B/LA	S75	S60	S45	S30	S15	0	15N	30N	45N	60N	75N
0	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2
10	44.7	63.9	70.4	74.1	76.8	79.1	81.4	84.1	87.8	94.3	113.5
20	0.0	41.4	55.8	63.5	69.0	73.8	78.5	84.0	91.7	106.1	0.0
30	0.0	12.0	39.4	51.9	60.6	67.9	75.3	83.9	96.5	123.8	0.0
40	0.0	0.0	19.4	38.6	50.9	61.2	71.5	83.8	103.0	0.0	0.0
50	0.0	0.0	−8.6	22.6	39.5	53.1	66.7	83.7	114.8	0.0	0.0
60	0.0	0.0	0.0	2.5	25.4	43.0	60.5	83.4	0.0	0.0	0.0
70	0.0	0.0	0.0	−23.1	8.2	29.9	51.7	82.9	0.0	0.0	0.0
80	0.0	0.0	0.0	−54.9	−12.1	13.4	39.0	81.7	0.0	0.0	0.0
90	0.0	0.0	0.0	−85.6	−33.0	−5.8	21.4	74.0	0.0	0.0	0.0
100	0.0	0.0	0.0	−93.3	−50.6	−25.0	0.5	43.3	0.0	0.0	0.0
110	0.0	0.0	0.0	−94.5	−63.3	−41.5	−19.8	11.5	0.0	0.0	0.0
120	0.0	0.0	0.0	−95.0	−72.1	−54.6	−37.0	−14.1	0.0	0.0	0.0
130	0.0	0.0	−126.4	−95.3	−78.3	−64.7	−51.1	−34.2	−3.0	0.0	0.0
140	0.0	0.0	−114.6	−95.4	−83.1	−72.8	−62.5	−50.2	−31.0	0.0	0.0
150	0.0	−135.4	−108.1	−95.5	−86.9	−79.5	−72.2	−63.5	−51.0	−23.6	0.0
160	0.0	−117.7	−103.3	−95.6	−90.1	−85.4	−80.6	−75.1	−67.4	−53.0	0.0
170	−125.1	−105.9	−99.4	−95.7	−93.0	−90.7	−88.4	−85.7	−82.0	−75.5	−56.3
180	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2	84.2

Mercator) and compute where the bearing tracks cross these selected latitude and longitude lines. In this program, northern latitudes and western longitudes are taken as positive. Southern latitudes and eastern longitudes are negative.

The program begins by asking you to provide the geographical coordinates of the origin (i.e., your location) in program **lines 90 and 100.** Then for a selected number of bearings, B, from this origin, it computes the maximum latitude, LV, which each bearing track reaches (the vertex) as well as the longitudes, NV, of these vertices. Great circle tracks for bearings, B, every 10 degrees from zero to 180 (**line 310**) were selected for computation. This gives complete

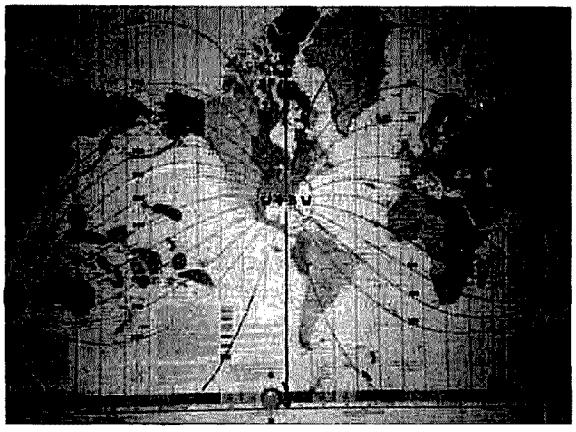


fig. 3. W4BW map, HO 5129, showing great circles from Tallahassee, Florida. This map was produced using the meridian intercept method of calculation.

360-degree coverage when you consider the great circle paths going in the reciprocal direction through the origin.

On the diagram, the great circle having a bearing, B, reaches the maximum latitude LV when angle C = 90 degrees. First we compute the co-latitude of the vertex CLV = (PI/2-LV) using the right spherical trigonometric identity (C = B in the program):

$SIN(CLV) = SIN((PI/2)-L1) \cdot SIN(B)$ (line **340**)
and then: $LV = (PI/2) - ARCSIN(CLV)$ (line **360**)
where $PI = 3.141593$ (line **60**)

The longitude of the vertex NV is then computed by first obtaining TV as defined on the diagram (the difference between the longitude of the origin, N1, and the longitude of the vertex NV) from the equation:

$TAN(TTV) = 1/((COS((PI/2)-L1)) \cdot TAN(B)$ (line **390**)
from which: $TV = ARCTAN(TTV)$ (line **400**)

The longitude of the Vertex NV is obtained from the definition of TV. Notice that there is a vertex both east (NVE) and west (NVW) of the origin:

$NVE = N1 - TV$ (line **420**)
$NVW = N1 + TV$ (line **440**)

The results of the computations for each selected bearing are stored in Array A(19,39) (line **290**) for subsequent print-out after all computations are made.

Next select the longitudes NS on which to determine the latitudes LX of the crossings of the (azimuthal) bearing tracks. In the program, NS was selected for every 30 degrees of longitude between −180 and 150 degrees (line **480**). Then compute LX for each bearing track at each of the selected longitudes NS as follows:

By definition $TH3 = NV - NS$ (line **500**)
then $TAN(LX) = COS(TH3) \cdot TAN(LV)$ (line **520**)
and $LX = ARCTAN(LX)$ (line **540**)

The longitudes of the bearing tracks NX at selected latitudes LS are found by using the following:

$COS(TH4) = TAN((PI/2) - LV)/TAN((PI/2-LS)$ (line **620**)
$TH4 = ARCCOS(TH4)$ (line **670**)
by definition $NX = NV + TH4$ (line **700**)

The selected latitudes LS in the program are every 15 degrees from −75 to 75 degrees (line **600**). A second table of bearing track crossings, NX, are also computed using the other vertex.

Program lines **810** through **1220** are the part of the program that prints out the answers. As an example, the results are given of computations made for the origin at the QTH of W4BW (latitude = 30.4N, longitude = 84.2W). **Table 1** is a list of latitudes and longitudes of the vertices for the W4BW location as well as TV (the degrees of longitude between the origin and the vertex) for the bearing angles selected. **Table 2** gives the latitudes of the great circle track crossings at the specified longitudes. **Tables 3** and **4** give the longitudes of the great circle track crossings on the specified latitudes. **Table 3** uses the longitude of the vertex east of the eastern vertex and **table 4** uses the longitudes of the western vertex. When these tables have been computed for your location as origin, the crossing points for each bearing given in the tables can be plotted on your map. When they are connected, the lines show the desired great circle paths emanating from your location.

conclusion

For those of you who use a calculator, remember that each bearing has two vertices, one east and one west of the origin. For azimuths from 0 to 90 degrees, the nearest vertex will lie east of the station, and for azimuths 90-180 degrees, the nearest vertex will lie west of the station. This applies to stations in the northern hemisphere. For stations in the southern hemisphere, the conditions will be reversed. When computing the nearest vertex for bearings of 90 and 270 degrees, note that it coincides with the station location.

Any map can be used as long as you can identify latitude and longitude lines in order to plot the intercepts. First connect the intercept points free-hand, using a soft pencil. Make the final lines with a water-base marking pen in any color you wish. Black is probably best.

Label the azimuths either by hand or with dry transfer lettering. For neatness, position the values verti-

cally on the Greenwich and 180th longitude meridians, for 0-180 and 180-360 degrees, respectively.

Readers who are too busy to do their own calculations can obtain an intercept computer readout for their locations from WØDW, for a nominal fee of $10.00, which covers the cost of paper and use of the computer. All you need to provide is the latitude and longitude of your location as accurately as possible.

For you avid DXers, nothing is easier or faster than glancing up at a map and instantly seeing the azimuth from your station to any location in the world. It's interesting to observe where your signal may travel with the beam heading on various azimuths. It also helps you understand some of the anomolies encountered with stations in the far corners of the earth when turning your beam makes little or no difference in signal strength. Lateral deviation of the signal path accounts for some of it, but in many cases it's the result of your signal reaching the area of the antipode regardless of the direction in which you point your beam.

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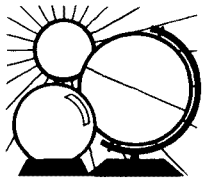
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DX FORECASTER

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summer propagation

Summertime QRN is this month's topic. The predominating noise source over omnipresent local noise is atmospherics, the cumulative propagated noise from some distance away. It isn't associated with weather storm fronts passing your QHT, but rather with air mass thunderstorms. These large high-pressure air masses cover several states and can linger for several days — even a week or more.

Some regions of the country have a greater number and variety of thunderstorms than others; in these areas, thunderstorm activity is measured by the number of "thunderstorm-days," or days in which at least one thunderstorm occurs. Areas with 100 thunderstorm-days or more are found in Florida and in the Rocky Mountains; the southern parts of Louisiana, Alabama, and Georgia endure 80 days. A band stretching across Nebraska to Ohio and then bending southward into South Carolina, and another reaching across New Mexico to Northern Texas, have 60 days, with the rest of the Mid-west experiencing 50 thunderstorm-days. Notice that most of the high numbers of thunderstorm-days occur in the southeast.

Because frontal weather activity is limited in these areas, the high numbers can be attributed mostly to air-mass thunderstorms. Moist air from the Gulf of Mexico moving northeast along the land serves as a prime thunderstorm generator.

Thunderstorm activity depends on how much moisture the air contains. If the air is dry, perhaps no thunderstorms will form at all. If the relative humidity is 50 percent or more, they may develop by noontime.

The process works as follows. First the sun heats the ground. As the heat from the ground rises, it warms the air above it and causes it to rise. As this heated air meets the colder air above, its moisture condenses, forming clouds. The clouds — some of which may be seized by the winds and carried into the jet stream to form the characteristic anvil-shaped top of a thunderstorm at 30 to 40,000 feet — continue to rise until their condensed moisture forms heavy drops heavy enough to fall as rain; some drops are taken further upward and freeze into hail. This fast up-and-down motion generates static electricity strong enough to cause the air to break down between a cloud and the earth or between one cloud and another. As the lightning stroke releases this energy, both sound and electromagnetic pulses are produced. Our receivers pick up the HF radio frequency pulse and we call it "noise."

After several days the air dries out or slowly moves on. However, during these days QRN may limit the usefulness of the low-frequency HF bands, enabling only local ragchewing and ruling out weak-signal DX reception.

How can you communicate with DX stations on these bands? Directional antennas may help if the thunderstorm activity is in the opposite direction from the DX stations. If you're using phased verticals you may be able to form a null in the right direction, or rotate one of the big beams to a noise null and then hear the DX. Receivers with effective noise blankers also spare your ears when the static crashes occur. When the going is really rough, most operators switch operating hours, giving up evening hours in favor of the pre-dawn hours. By this time the thunderstorms have lost their heat and dissipated to the east, locally, and are weakening on DX paths to the west as the sun moves across Asia and the Middle East. In addition, the sun in the morning hasn't yet warmed the air enough for the day's thunderstorms to occur. This is a cool, comfortable time of the day to be up and around.

last-minute forecast

The higher-frequency HF bands, 10 through 30 meters, are expected to be best for long-skip the first week and the last few days of August. Expect sporadic-E short-skip to be the best of the summer season on 6 meters and lower in frequency a good many hours of many days of the month. The lower bands aren't expected to be very good in the evenings because of the air-mass thunderstorm build-up. These bands can provide some good DX contacts across the continent or into Europe for a short period in very early morning on those days during the third week, when the solar flux is at its lowest. Geomagnetic disturbances will occur less frequently, with the greatest possibility during the third week.

For the VHF/UHF enthusiast the moon's perigee will occur on the 16th, with full moon on the 27th. The Perseids meteor shower will occur from the 10th to 14th, with a maximum rate of better than 50 meteors per hour expected on the 11th and 12th. This is an excellent shower to work with.

band-by-band summary

Six-meter sporadic E short-skip conditions will occur for a half hour to several hours around local noon on some days for this last good month of this summer's E_s season. Expect about 1000 miles per hop.

Ten, twelve, and fifteen meters will have a quite few short-skip E_s openings and some long-skip openings to southern areas of the world during daylight hours. Fifteen meters will be best for only an hour or two as the maximum usable frequency decreases during the late afternoon.

Twenty, thirty, and forty meters will be useful for distance communications to most eastern, western, and northern areas of the world during daytime and into the evening almost every day,

GMT	PDT	WESTERN USA N	NE	E	SE	S	SW	W	NW	MDT	MID USA N	NE	E	SE	S	SW	W	NW	CDT	EDT	EASTERN USA N	NE	E	SE	S	SW	W	NW
0000	5:00	20	20	20	10	15	10	10	15	6:00	20	20	20	10	20	10	10	20*	7:00	8:00	20	30	20	12	20	10	10	20
0100	6:00	20	*30*	20	12	15	10	10	15	7:00	20	*30*	20	12	20	10	10	20	8:00	9:00	20	30	20	12	20	10	10	20
0200	7:00	20	*30*	*20*	12	20*	10	10	15	8:00	20	*30*	*20*	15	20	10	10	20	9:00	10:00	20	30	*20*	15	20	10	10	20
0300	8:00	20	20	*20*	15	20	10	10	15	9:00	20	*30*	*20*	20	20	10	10	20	10:00	11:00	30	*40*	*20*	20	20	10	12	20
0400	9:00	20	20	*30**	20	20	10	10	20	10:00	30	30	*30**	20	30	10	10	20	11:00	12:00	30	*40*	*30**	20	30	12	15	30
0500	10:00	20	20	*20*	20	30	10	10	20	11:00	30	20	*20*	20	30	12	10	30	12:00	1:00	*30*	*40**	*20*	20	30	15	15	30
0600	11:00	20	20	20	20	30	12	12	20	12:00	30	20	20	*20*	30	15	12	30	1:00	2:00	*30*	*40*	20	*20*	30	15	20	30
0700	12:00	30	20	20	20	30	12	12	*20*	1:00	*30*	*30*	*20*	*20*	30	15	15	30	2:00	3:00	30	*40*	*20*	*20*	40	20	20	*30*
0800	1:00	30	*30*	20	*20*	30	15	15	20	2:00	*30*	*30*	*20*	*20*	40	20	20	*30*	3:00	4:00	20	30	*20*	*20*	40	20	20	*30*
0900	2:00	*30*	*30*	20	*20*	30	20	20	30	3:00	*40**	*30*	20	*20*	40	20	20	*30*	4:00	5:00	20	20	20	*20*	*40*	20	*20*	30
1000	3:00	*30*	*30*	20	*30**	*40*	20	20	30	4:00	30	20	15	*20*	*40*	20	20	30	5:00	6:00	20	20	15	*20*	*40*	*20*	*20*	30
1100	4:00	*40**	20	20	*20*	*40*	20	20	*30*	5:00	20	20	15	*20*	*40*	*20*	20	30	6:00	7:00	20	20	15	*20*	*40*	*20*	*20*	30
1200	5:00	20	20	15	*20*	*40*	*20*	20	*30*	6:00	20	20	12	15	*40*	*20*	20	30	7:00	8:00	20	20	12	15	*40*	20	*20*	30
1300	6:00	20	20	12	20	*40*	*20*	*20*	*30*	7:00	20	20	12	15	*40*	*20*	*20*	30	8:00	9:00	20	20	12	15	*40*	*20*	*20*	30
1400	7:00	20	20	12	15	*40*	*20*	*20*	*30*	8:00	20	20	10	12	*40*	*20*	*20*	*30*	9:00	10:00	20	20	10	12	*40*	*20*	*20*	30
1500	8:00	20	20*	10	12	*40*	*20*	*20*	*30*	9:00	20	15	10	12	*40*	*20*	*20*	*30*	10:00	11:00	20	20	10	12	30	*20*	*30*	*30*
1000	9:00	30	15	10	12	30	*20*	*30*	*30*	10:00	30	15	10	12	30	*20*	*20*	*30*	11:00	12:00	20	20	10	10	20	*30**	*30*	*30*
1700	10:00	30	15	10	12	20	20	*30*	*30*	11:00	30	20*	10	10	20	20	*30*	*30*	12:00	1:00	30	20	10	10	20	20	*30*	*40**
1800	11:00	*30**	20*	10	10	20	20*	*30*	20	12:00	*30*	20	10	10	20	15	*30*	20	1:00	2:00	30	20	10	10	20	15	*30*	20
1800	12:00	*30*	20	12	10	15	15	20	20	1:00	*30*	20	12	10	20*	12	*30*	20	2:00	3:00	*30*	20	12	10	20	12	20	20
2000	1:00	*40**	20	12	10	15	12	12	20	2:00	*30*	20	15	10	15	12	20	20	3:00	4:00	*30*	20	12	10	20*	12	12	20
2100	2:00	20	20	15	10	15	12	12	20	3:00	20	20	15	10	15	10	15	20	4:00	5:00	*30*	20	15	10	15	10	10	20
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2300	4:00	20	20	20	10	15	10	10	15	5:00	20	20	20	10	15	10	10	15	6:00	7:00	20	30*	20	10	20	10	10	20*
AUGUST		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN			ASIA FAR EAST	EUROPE	S. AFRICA	CARIBBEAN S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.
*Look at next higher band for possible openings.

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either long-skip to 2000 miles per hop or short-skip E_s with 1000-mile hops. The period of daylight is still relatively long, but will be noticeably shorter by the end of the month.

Thirty, forty, eighty, and one-sixty meters are all good for nighttime DX even though the background noise is severe in the evenings. The direction of the openings will rotate around from the east, to the south, and then westward toward the morning. Sporadic E layer propagation may be of help in the early evening toward the east and south to override thunderstorm noise. Try those early morning hours for communication paths to the west and monitor WWV or WWVH on 2.5 and 5 MHz.

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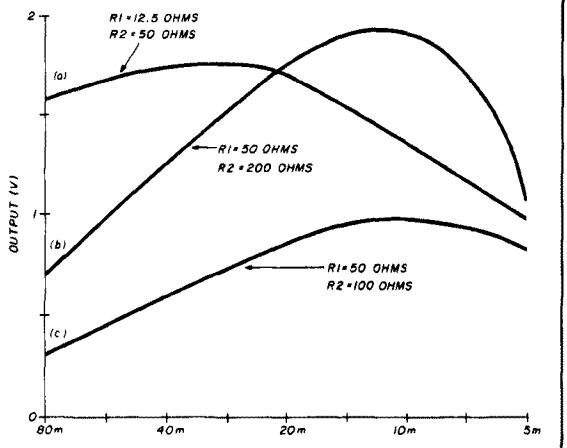


fig. 1A. Isolated winding line transformer response curves: (a) unbalanced configuration, lower impedance level; (b) unbalanced configuration, higher impedance level; (c) balanced configuration.

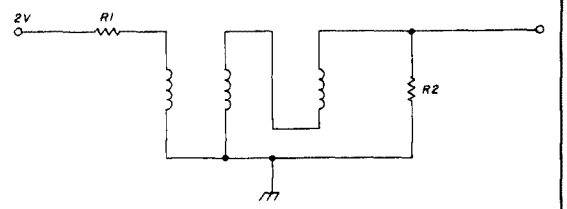


fig. 1B. Isolated winding line transformer, unbalanced configuration.

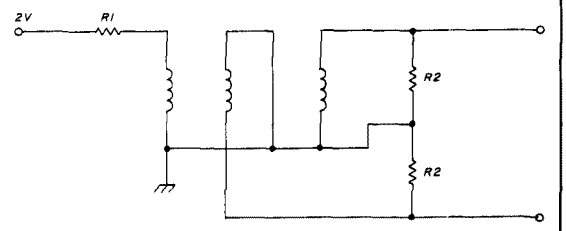


fig. 1C. Isolated winding line transformer, balanced configuration.

HF, VHF or UHF performance without ferrite cores

Line transformers are usually associated with antenna matching or VHF/UHF stripline applications. This article, however, describes general-purpose transformers made from wire. They're simple to make and can serve in many applications normally reserved for ferrite core transformers. All the examples considered use tightly twisted trifilar enamel or Formvar-coated No. 22 wire, with approximately 1/8 inch between twists. Because each wire is comparable to one winding on a conventional trifilar transformer, these transformers can be operated in all the normal trifilar configurations. Relatively broadband, their center frequency response is determined by the wire length — several inches for UHF operation and several feet for the lower HF frequencies. Although the trifilar twisted wire could be wound on a small diameter form, little difference in performance would be noticed between it and an uncoiled version.

Measurements have been made on a 17-inch long trifilar wire transformer coiled on a 5/16-inch diameter, 3/4-inch long form. Test results and configurations are shown in **figs. 1, 2** and **3.** Resonance is influenced by both the configuration and loading. The balanced configuration with the inductors in parallel **(fig. 1)** has the highest frequency response, while the 3:1 balun with the windings in series **(fig. 2)** has the lowest frequency response. Inherently a low-impedance device, its lower termination resistance minimizes reflections and results in a wider bandpass. This is illustrated in

By John C. Reed, W6IOJ, 770 La Buena Tierra, Santa Barbara, California 93111

the 2:1 balun (**fig. 3**), where the 200-ohm termination shows a dominant resonant condition as compared to that of the 50-ohm termination. This 17-inch transformer is used in a test equipment balanced modulator application using the circuit shown in **fig. 4**. It provides 100 milliwatts at 21 MHz, with the balanced transformer isolating the modulated output from the input excitation by over 40dB.

My primary application for the transformers has been in double-balanced mixers; an example is shown in **fig. 5**. The UHF transformer, with only a 3-inch trifilar wire, isn't coiled, but instead mounted in a convenient loop arrangement forming an interface with the circuit. The mixer loss is −6 to −8 dB, and the poorest port-to-port isolation, between the local oscillator and output, is over 20 dB.

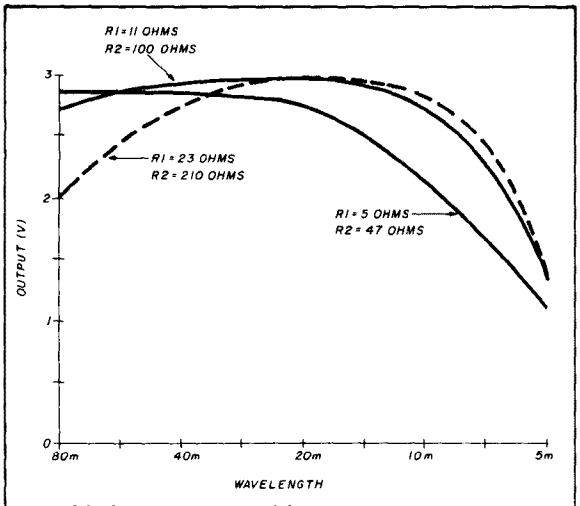


fig. 2A. 3:1 (voltage ratio) balun: frequency response.

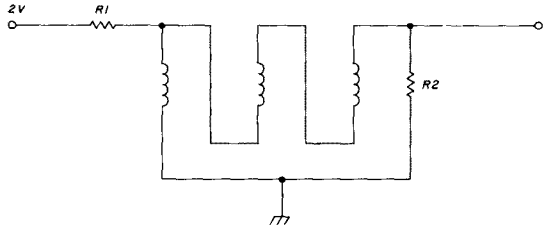


fig. 2B. 3:1 (voltage ratio) balun: circuit configuration.

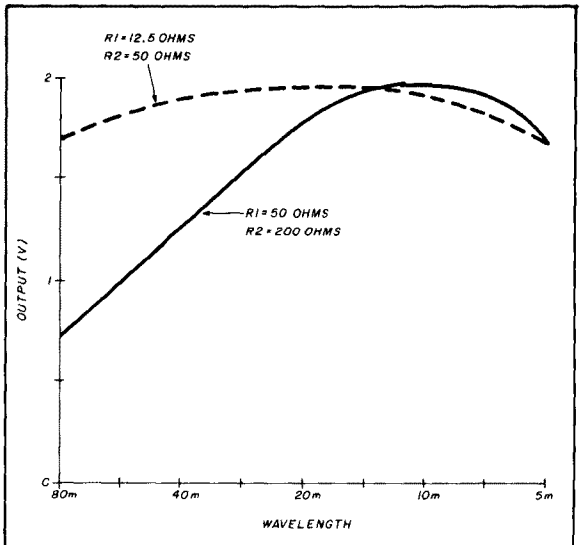


fig. 3A. (voltage ratio) balun response with two different set of termination conditions.

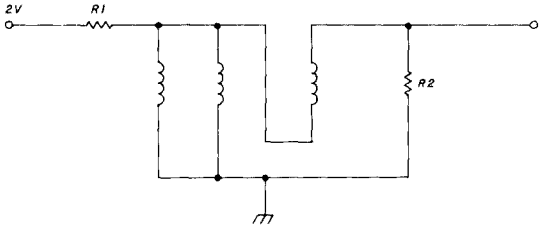


fig. 3B. 2:1 (voltage ratio) balun: circuit configuration.

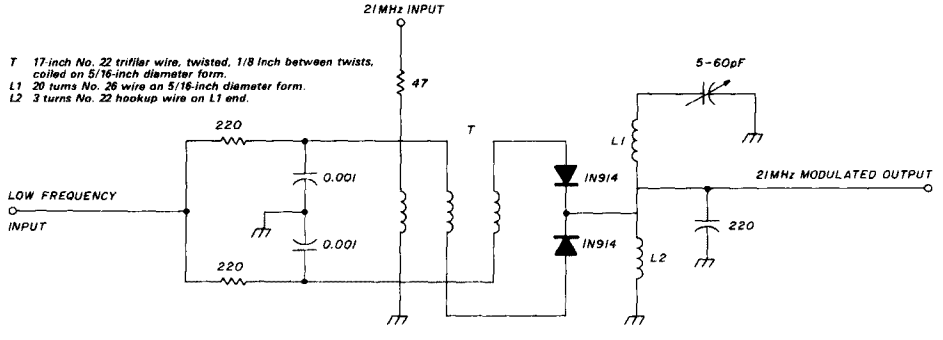
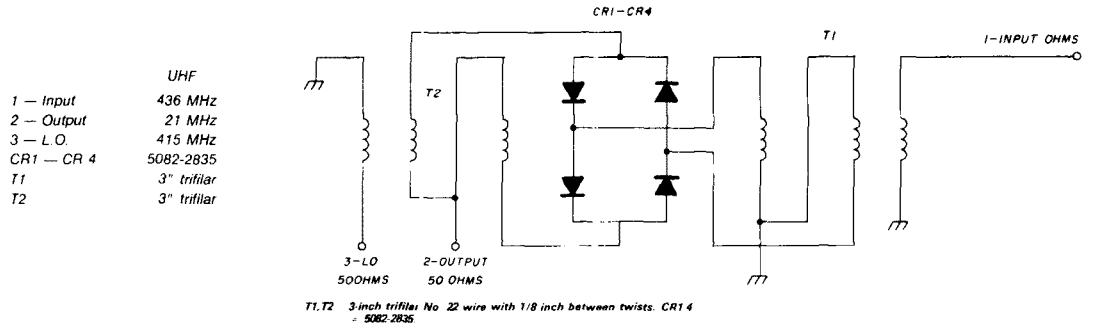


fig. 4. Balanced modulator circuit using trifilar wound line transformer.



	UHF
1 — Input	436 MHz
2 — Output	21 MHz
3 — L.O.	415 MHz
CR1 — CR 4	5082-2835
T1	3" trifilar
T2	3" trifilar

fig. 5. Double balanced mixer.

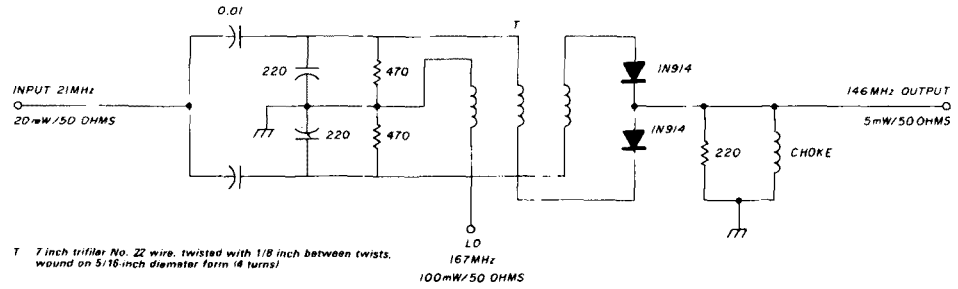


fig. 6. Single balanced mixer provides 2-meter output with 21 MHz input and 167 MHz LO.

I have also used the trifilar wire transformer in a 2-meter single balanced mixer (**fig. 6**).

This experimental effort was initiated after questionable results were obtained from homebrew UHF trifilar ferrite core transformers. The wire transformers solved my problems with consistent and satisfactory results.

The technique is simple enough to try out in a few minutes. Using easily obtainable wire (I used Radio Shack No. 276-1345), stretch the three wires between a solid tie point and a variable-speed drill, turning the drill for the desired 1/8-inch twist spacing. I quickly check the resonant frequency of a particular wire-length/configuration by simply coupling the input to a grid-dip meter, detecting the transformer output with a 1N34A diode peak detector.

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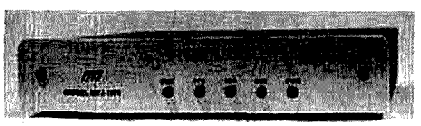
NEW products

new TNC from MFJ breaks price barrier

MFJ Enterprises, Inc. of Starkville, Mississippi, has released its new affordable packet radio controller.

At $129.95 (plus $6 shipping), the MFJ-1270 breaks a price barrier to make packet radio more affordable by bringing together efficient manufacturing techniques and TAPR'S Tucson Amateur Packet Radio) leading-edge technology.

The MFJ-1270 is a nearly identical clone of the widely acclaimed TAPR TNC 2, with identical software and hardware. Neatly packaged in a new cabinet, the unit includes a TTL serial port for extra versatility.



All that's needed to operate packet radio is a rig and any home computer with an RS-232 serial port and a terminal program.

For Commodore 64, 128, or VIC 20 users there's an optional Starter Pack for $19.95; this puts them on the air immediately and includes interfacing cable, terminal soltware on tape or disk, and complete instructions — everything the users need to get on packet radio. The MFJ-1282 is the disk version and the MFJ-1283 is the tape version. Both are priced at $19.95 each.

Unlike machine-specific TNCs, the general-purpose MFJ-1270 won't become obsolete because the user changes computer or because packet radio standards change. The MFJ-1270 can be used with *any* computer that has an RS-232 serial port and an appropriate terminal program. If packet radio standards change, software updates will be made available as TAPR releases them.

Speeds in excess of 56 Kilobauds are possible with a suitable external modem. These faster speeds would be very difficult, if not impossible, with any machine-specific TNC or a TNC without hardware HDLC. The MFJ-1270 can be used as an inexpensive digipeater.

The MFJ TNC uses the latest AX.25 Version 2.0 software. As few as three commands can be used to make contacts; 82 software commands are featured. The "MHEARD" command lists stations and the time heard. Up to four station roundtables are possible.

Hardware features include HDLC for full duplex, Z80A CPU, and a 16K EPROM, expandable to 32K. 16K RAM (with battery backup) and true data carrier detect are included. Terminals up to 9600 baud may be used. A tuning indicator port for HF and satellite operation is featured, and LEDs indicate connection, packet tone received, packet acknowledgement, power on, and activation of transmitter.

The MFJ-1270 Packet Radio TNC 2 and MFJ-1282/MFJ-1283 Starter Packs come with a double guarantee. They can be ordered from MFJ for a 30-day trial period. If they don't satisfy your needs, return them for a full refund, less shipping. MFJ's full one-year unconditional guarantee covers all products.

To order or request additional information, contact MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, Mississippi 39762.

Circle #301 on Reader Service Card.

digital voice recorder

Advanced Computer Controls, Inc., has announced its new Digital Voice Recorder, which provides remote audio record and playback capability for voice repeaters. The DVR allows the repeater owner to remotely record any of the repeater's programmable messages — IDs, tail messages, emergency autodial responses, alarms, courtesy tones, etc. In addition, its touchtone-activated voice mailbox lets users leave recorded messages for others.

The DVR's direct digital recording technique offers a high-quality audio mode as well as two additional quality level modes designed to conserve memory and extend recording time.



The DVR uses up to 32 inexpensive 64K or 256K dynamic RAM chips and is available with one, two, or three independent record/playback channels, which means that one DVR can service up to three repeaters at one site, or two repeaters and the phone line.

The DVR is available from $849 (manual alone, $5 postpaid). For additional information, contact Advanced Computer Controls, Inc., 10816 Northridge Square, Cupertino, California 95014.

Circle #302 on Reader Service Card.

Butterworth design program for IBM PC and compatibles

Etron RF Enterprises has released the third in their series of PC-DOS programs for designing RF circuitry. RF Notes Number 3, Volume 1 covers low-pass, high-pass, bandpass and band-reject Butterworth filters. Each of the four programs outputs both graphical response curves and schematic diagrams.

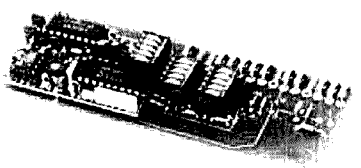
Written in BASIC A, the programs are available in both monochrome and color versions for the IBM PC, PC XT, PCjr and most compatibles. Each requires 128K of memory and the appropriate graphics display card.

Further information is available from Etron RF Enterprises, P.O. Box 4042, Diamond Bar, California 91765. (Also available from Ham Radio's Bookstore, Greenville, NH 03048, for $84.95 plus $3.50 shipping and handling.)

Circle #303 on Reader Service Card.

plug-in encoder-decoder for RCA/TACTEC TAC-200

Communications Specialists, Inc. of Orange, California has just introduced a direct plug-in encoder-decoder for the RCA/TACTEC TAC-200 and VEETAC mobiles. The TS-32TAC, based on the proven TS-32, is available in two versions; the TS-32TAC-1 will encode and decode one tone only, while the TS-32TAC-2 will either encode and decode two tones, or encode one tone and decode a different tone.



Both versions of the TS-32TAC plug directly into the host radio without any modifications to the tone board or the radio, to allow for economical installation and removal.

The TS-32TAC is available for immediate shipment and comes complete with a one-year warranty. Prices are $78.26 for the TS-32TAC-1 single tone unit and $82.92 for the TS-32TAC-2 *two-tone board*.

A catalog is available on request. For information, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665.

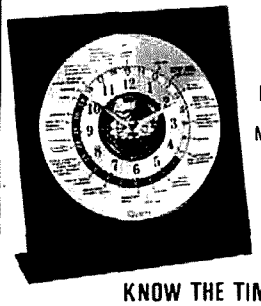
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ham upgrade cassettes and books

Gordon West's Radio School has produced a four-cassette stereo theory course entitled "Tapes'N Books Theory Course." Each course contains four long-play, stereo theory tapes plus a fully illustrated textbook and the ARRL's *FCC Rule Book*. Theory cassette courses are available for all grades of amateur radio licensing.

The course contains the questions and answers to be found on any Amateur Radio examination, as well as the "sounds" of Amateur Radio. You can actually hear the difference between a properly adjusted speech processor and the effects of overmodulation, for example. You can hear the difference between AMTOR, ASCII, and BAUDOT. You'll tune into actual ham radio transmissions, and actually hear the

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9258	RG-8X	100	37	37

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PL-259	UHF Male cable end silver	1.65
PL-258	UHF Barrel connector	2.00
UG-175	Reducer for RG-58 cable	.45
UG-176	Reducer for RG-59 Mini-X	.45
UG-88	BNC Plug for RG-58	2.00
UG-260B	BNC Plug for RG-59	2.10
UG-260B	BNC Plug for Mini-X fitted	3.60
UG-625B	BNC panel receptacle	1.45

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All courses are $19.95 plus $5.00 for first class, same-day, postage, handling, and mailing.

For more information, contact Gordon West Radio School, 2414 College Drive, Costa Mesa, California 92626.

Circle #305 on Reader Service Card.

repair or build antennas — with ease

At last there's a brazing rod for aluminum with a low melting point of only 732 degrees. All that's needed for easy antenna construction or repair is a hand-held propane torch, a good wire brush, and a supply of M.S.S. Wonder Rods. Only three simple steps are necessary: (1) clean the area to be welded with wire brush; (2) heat the area to be welded until the rod flows, scrubbing with the wire brush; and (3) heat the area again until the rod flows, fill it in, and let it cool.

In building or repairing antennas, you can eliminate high-resistance joints and the need for clamps by welding. Antenna elements damaged by wind or sleet can be straightened; kinks or cracks can simply be filled in.

Other uses for M.S.S. Wonder Rods include welding copper to aluminum (as in a matching network), repairing or plugging holes in aluminum panels, repairing holes in aluminum boats, or repairing most of the parts of your alternator. Other metals you can weld with Wonder Rods include copper, brass, galvanized steel, and white metal.

A kit containing Wonder Rods and a stainless steel brush is available for $19.95, which includes tax and postage. For details, contact M.S.S. Wonder Rods, N3401 Castle Road, Medford, Wisconsin 54451.

Circle #306 on Reader Service Card.

"clamp-on" RF choke suppresses interference

EMC Datacare's D910 series of clip-on chokes are said to greatly reduce interference when they're clamped onto interconnecting cables of various types of electronic systems. For example, installing them between the various parts – external drives, monitors, and printers – of a computer system provides significant reduction of any computer-caused RFI problems.



The chokes clamp onto cables up to 0.4 inch in diameter. A professional kit (D918), designed to protect up to eight cables, is available.

For details, contact EMC Datacare Ltd., Powell Court, Luton, Bedfordshire, LU1 3JJ, England.

Circle #307 on Reader Service Card.

packet switch allows instant change from 1200 to 300 baud operation

Amateur Packet Alaska has announced the availability of its TNC VHF/HF Switch Kit. It's all-CMOS logic assembly quickly changes critical filter and timing components in the TAPR, AEA, MFJ, Heath and Pac-com TNCs to optimize the on-board modem for HF or VHF use.

The easy-to-build kit installs easily inside the TNC's case. Complete step-by-step instructions are included; construction should be completed in just one hour.

The APA VHF/HF Switch is available for $30 which includes airmail postage. As an all volunteer organization, APA is unable to accept telephone or credit card orders. But you can write to them for more information – or to place an order – at AX.25 Communications Trail, Ester, Alaska 99725.

Circle #308 on Reader Service Card.

heat-powered "battery"

A unique electrical generator has been developed and is being marketed by Ovonic Thermoelectric Company, a division of Energy Conversion Devices, Inc. This compact, lightweight, silent, solid-state generator obtains its energy from any convenient source of heat and produces up to 6 watts of DC power at 9 volts and up to 5 watts at 6 or 12 volts. According to Ovonic, it can be thought of as a heat-powered battery that never needs recharging.

Designated CSG (Compact Silent Generator), it can serve either as a primary source of power or as a backup or emergency power source, and can power portable televisions, radios, lamps, fans, pumps or other light electrical loads. It can also be used with an accessory to recharge batteries.

Campers can operate the CSG on camp stove or set it on a grate over a campfire. Boaters and operators of recreational vehicles can use it on the galley stove. In homes and cottages

NEW products

equipped with wood stoves, the CSG can be placed directly on the stove to produce electricity. In homes it can also provide emergency power from a gas stove or can of Sterno™.

For information, contact Energy Conversion Devices, 1864 Northwood, Troy, Michigan 48084.

Circle #309 on Reader Service Card.

Monitoring Times merges with *International Radio*

Bob Grove, Editor of *Monitoring Times*, and Larry Miller, Editor of *International Radio* (formerly *Shortwave Guide*), have announced plans to combine their publications into a single operation.

Monitoring Times, now in its fifth year of publication, has grown steadily, earning a worldwide reputation for accuracy, timely reporting, and journalistic integrity. *International Radio* has similarly earned a reputation for in-depth interviews of broadcasting personalities, up-to-the-minute time and frequency schedules, and program details.

The expanded 60-page tabloid will be the largest periodical in the monitoring industry, featuring authoritative content and improved printing from a new laser printer on heavier paper.

A year's subscription to the new *Monitoring Times* will be only $14. Discounts are available for two- and three-year subscriptions, quantity orders for resale, and educational institutions.

For more information on domestic or foreign subscriptions, contact *Monitoring Times* at 140 Dog Branch Road, P.O. Box 98, Brasstown, North Carolina 28902.

Circle #310 on Reader Service Card.

160-meter vertical

Oak Hills Research has announced its new DX-60V 160-meter vertical antenna, a folded design that measures 160 feet overall, with 120 feet of wire. The DX-60V can be installed in a tree, on a tower, in either a slanted or vertical position. It will be handle the full legal limit and has a typical bandwidth of 110 kHz between 2:1 SWR points. High-impact polystrene coil frames and spreaders are used to prevent breakage in high winds and cold weather. Tapping the loading coil at two points and switching between them will give band coverage. In order to achieve maximum performance, Oak Hills warns that it's imperative to have a ground system. Several suggestions are provided in the maufacturer's literature.

For more details, contact Oak Hills Research, 4961 North Douglas Road, Luther, Michigan 49656.

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LOW FREQUENCY Knight S/R receiver 200 kHz-30 MHz $55. RCA 308-B field intensity meter 120 Hz-1.8 MHz $150. HP-310A wave analyzer receiver 1 kHz-1.5 MHz $175. HP0-100D L/F standard 10 Hz-1 MHz has 2" scope $85. HP-202C L/F oscillator 1 Hz-100 kHz $45. Trade for FT-757GX or TS-430S. K6KZT, 2255 Alexander, Los Osos, CA 93402.

BEAM HEADING CHART, 10 page report in binder with 9 data fields calculated from your exact QTH to over 540 DX locations. $9.95 from John Daley, KB6JGH, PO Box 4794, San Jose, CA 95150.

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THE GOOD SAM HAMS invite RV operators to check in the Good Sam Ham net 14.240 *Sundays 1900Z also 3.880 Tuesdays at 2359Z. Net control N5BDN, Clarksville, IN*

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COMING EVENTS

Activities — "Places to go . . ."

THANK YOU! TCRA (Tri-County Radio Association) thanks all our customers and vendors who made our May 4, Stirling, NJ, Hamfest a big success. We look forward to seeing you next year.

ALABAMA: The Huntsville Hamfest, Alabama's largest, Saturday and Sunday, August 16 *and 17, Von Braun Civic Center,* Huntsville. No admission charge. Exhibits, forums, ARRL booth. Also non-ham activities. Air-conditioned indoor flea market. Reserved tables $8/day. Walk in CAVEC exams given at the Huntsville HS cafeteria beginning 9 AM Saturday. Tours of the Alabama Space & Rocket Center available. A limited number of camp sites with hookups available at the VBCC. Talk in on 34/94. For more information: Huntsville Hamfest, 2804 S. Memorial Parkway, Huntsville, AL 35801.

OHIO: The Lancaster and Fairfield County Amateur Radio Club's annual Hamfest, Sunday, August 10, Fairfield County Fairgrounds, Lancaster. 8 AM to 4PM. Admission $3.00 advance or $4.00 at the door. Limited tables at $4.00 each advance; $5.00 at door or bring your own. Table space $3.00 advance or $4.00 at door. Plenty of parking. Refreshments available all day. For more information write Box 3, Lancaster, Ohio 43130.

ILLINOIS: The Hamfesters Radio Club is sponsoring their 52nd Hamfest, Sunday, August 10, Santa Fe Park, 91st and Wolf Road, Willow Springs (near Chicago). Gates open 6 AM. Free parking. Vendors, ARRL and FCC tables. Refreshments available at reasonable prices. Tickets $3.00 donation advance. $4.00 donation at gate. For advance tickets send check and SASE to Hamfesters Radio Club, PO Box *42792, Chicago, IL 60642.* For *general* information call (312) 598-4802.

OHIO: The 12th annual Hall of Fame Hamfest, August 10, presented by the Tusco ARC, W8ZX and the Canton ARC, W8AL. Nimishillen Grange, 6461 Easton Street, Louisville. Registration $2.50 advance; $3.00 at the gate. Reserved tables available $5.00 each. Deadline 8/1/86. Large flea market, forums and plenty of *good food. Mobile check in on 146.52/52 and 147.72/12. For in*formation: Tusco ARC, PO Box 725, New Philadelphia, OH 44663.

NEW YORK: Electronics Extravaganza sponsored by the Putnam Emergency Amateur Repeater League, Saturday, August 16, J.F. Kennedy Elementary School, Brewster. 9 AM to 4 PM. Admission $3.00. Children under 12 free with adult. Food and refreshments available. Indoor tables, with one admission, $10.00. Tailgating $5.00. FCC *exams.* Limited walkins. Talk in on 144.535/145.135. For table reservations or information: R. Dillon, N2EFA, RFD 7, Noel Court, Brewster, NY 10509.

WYOMING: ARRL Rocky Mountain Division Convention and 5th annual WIMU Hamfest, August 1, 2 and 3, Virginian Motel, Jackson Hole. Offering the scenic beauty of Yellowstone and Teton National Parks. Great fishing, horseback riding, raft rides on the Snake *River. Fun for the entire family!* For more information: George Siegel, 130 E. 17th Street, Idaho Falls, ID 83401. Telephone (208) 523-7433.

PENNSYLVANIA: W3PIE, the Uniontown Amateur Radio Club's 37th annual Gabfest, Saturday, September 6, Club Grounds, Old Pittsburgh Road, Uniontown, 50 miles south of Pittsburgh. Free parking, free coffee, free Swap & Shop setup with registration at $3.00 each or 2/$5.00. Talk in on 147.645-.045 and 144.57-.17. For further information: UARC Gabfest Committee c/o John T. Cermak, WB3DOD, PO Box 43, Republic, PA 15475. (412) 246-2870.

PENNSYLVANIA: The Central PA Repeater Association's annual Electronic Exhibit/Ham/Computerfest, August 24, Penn Harris Inn and Convention Center Rts 11 and 15, Camp Hill. Gates open 8 AM. Admission $4.00. XYL's and children 12 and under free. Indoor air-conditioned exhibit area. Large outdoor tailgating area. Various seminars. Amateur license exams starting 9 AM. Food and refreshments available. Talk in on 145.47 and 146.52 simplex. For information and reservations: Paul McDonnell, N3BKI, 1207 Apple Drive, Mechanicsburg, PA 17055. (717) 697-1880 (12 noon to 8 PM)

INDIANA: The 9th annual Bloomington Hamfest, Sunday, August 31, 2335 Vernal Pike, the 147.78/.18 repeater site. 8 AM to 2 PM. Admission $2.00. Food concession. No charge for setups but bring own tables. FCC VE exams contact K9PS for details. For Hamfest information please SASE to Bob Myers, K9KTH, 306 S. Fairview St., Bloomington, IN 47401 or call (812) 332-1105.

MAINE: ARRL sanctioned Windsor Hamfest, sponsored by the Augusta Emergency Amateur Radio Unit, September 6, Windsor Fairgrounds. Flea market, programs, speakers, commercial distributors and the traditional Saturday bean and casserole supper. Gate donation $1.00. Camping $3.00 per night or $5.00 for two nights. Talk in on 146.22/82 repeater. For more information: Dot, W1TGY and Phil, W1JTH, Young, 47 Longwood Avenue, Augusta, ME 04330. (207) 622-1385.

INDIANA: The Porter County Amateur Radio Club's Hamfest, August 3, 49'er Drive-In Theater, State Road 49, Valparaiso. Gates open 8 AM. Admission $3.00 at gate. Talk in on 147.96/.36 and .52 simplex. For more information: PCARC, PO Box 1782, Valparaiso, IN 46382.

NEW JERSEY: The Ramapo Mountain Amateur Radio Club's 10th annual Flea Market, August 23, Oakland American Legion Hall, 65 Oak Street, Oakland. 20 miles from GW bridge. Indoor tables $6.50. Tailgating $3.00. Admission $1.00. Non-ham family members free. Talk in on 147.49/146.49 and 52. For information: Tom Risseeuw, N2AAZ, 63 Page Drive, Oakland, NJ 07436. Tel. 337-8389 after 6 PM.

INDIANA: The 7th annual Grant County Amateur Radio Club Hamfest, Sunday, August 10, 4-H Fairgrounds, Marion. Doors open 8 AM. Refreshments, free parking license exams. Donation $2.00 advance; $3.00 gate. For information/tickets SASE to: WB9EAP, Brooks Clark, 2202 South Boots Street, Marion, IN 46953.

MISSOURI: The St. Charles ARC will sponsor Hamfest '86, Sunday, August 24, Blanchette Park (New Site), St. Charles. 7 AM to 2:30 PM. Free admission and parking. Forums and FCC license ex-

ams at 10 AM. $2.00 donation for tailgate flea market. Dealers welcome. Talk in on 146.07/67 repeater and 146.52 simplex. Contact: Eric Koch, NF0Q, 2805 Westminster, St Charles, MO 63301. (314) 946-0948.

MINNESOTA: The St. Cloud Amateur Radio Club Hamfest, August 10, Lake George, St. Cloud. Displays, demonstrations and trades. Ticket donation $3.00. Extra ticket $2.00. Snack counter. Talk in 34/94 primary, 615/015 secondary. Contact: SCARC, Box 141, St. Cloud, MN 56302.

TENNESSEE: The Lebanon Hamfest sponsored by the Short Mountain Repeater Club, August 31, Cedars of Lebanon State Park, US 231, Lebanon. Outdoor facilities only. Exhibitors bring your own tables. Food and drinks available. Talk in on 146.31/146.91. For more information, Mary Alice Fanning, KA4GSB, 4936 Danby Drive, Nashville, TN 37211.

MICHIGAN: September 14, L'Anse Crease ARC will have its 14th annual Swap and Shop, L'Anse Creuse High School, Mt. Clemens. 0800-1500. Admission $1.00 advance, $3.00 at the door. Plenty of food and parking. Trunk sales $4.00/space. Inside tables $8.00 each. For Tickets and table reservations SASE to Maurice Schetecatte, N8CEO, 15835 Touraine Ct., Mt. Clemens, MI 48044. (313) 286-1843.

MASSACHUSETTS: The Chelsea Civil Defense will sponsor Amateur Radio evening classes at Chelsea High School starting September 11, 1986 for those wishing to obtain a Novice, Tech/General license. There will be a minimal fee for materials cost only. Share in a great hobby! For more information, Frank Masucci, 136 Grove Street, Chelsea, MA 02150. Please include your phone number.

VIRGINIA/WEST VIRGINIA: The Bluefield Hamfest, sponsored by the East River ARC, Sunday, Augusta 24, Brushlord Armory, 1 mile north of Bluefield, WV, US 52. Admission $4.00. Children under 12 free. Large indoor flea market. Amateur radio, computer, satellite TV and other specialty dealers. Walk-in Amateur Radio exams 9 AM only. $4.00 fee for all exams except Novice. Talk in on 144.89/145.49 and 146.52. Paved parking, food on site, other activities. For information: Jim Perdue, KC8NG, Rt. 5, Box 457, Bluefield, WV 24701.

MICHIGAN: The Five County annual Swap-N-Shop, August 24, Saginaw Civic Center. Doors open 8 AM, 6 AM for dealers. Advance tickets $2.00, $3.00 at the door. Tables $7.50/8' tables. Covered trunk sales area $3.00 per car. Talk in on 147.84/24 and 146.52. For information call Don (517) 893-3475. For advance tickets and table reservations: Five County Swap-n-Shop, PO Box 1783, Saginaw, MI 48605-1783.

OHIO: The Union County Amateur Radio Club's 10th annual Marysville Hamfest and Auction, Sunday, August 24, Marysville Fairground. Gates open 6 PM Saturday evening for free overnight camping, free entertainment Saturday night. Auction Sunday afternoon. Large flea market area. $1.00/10' space. Power $2.00 extra. Advance tickets $3.00. At the gate $4.00. For information please SASE to: Gene Kirby, W8BJN, 13613 US 36, Marysville, OH 43040. (513) 644-0468.

PENNSYLVANIA: The Mid-Atlantic Amateur Radio Club's annual Hamfest, Sunday, August 10, Bucks County Drive-in Theater, Rt. 611, Warrington. 8 AM to 3 PM. Rain or shine. Admission $3.00 plus $2.00 for each tailgate space. Bring your own tables. Talk in on WB3JOE/R, 147.66/.06 or 146.52. For information: MARC*PO Box 352, Villanova, PA 19085 or call Bob Josuweit, WA3PZO (215) 449-9727.

INDIANA: The WA9SNT Amateur Radio Club will hold its annual Swaplest, August 9, ITT Technical Institute, 9511 Angola Court, Indianapolis. 8 AM to 4 PM. General admission $2.00. Students $1.00. Flea market space $1.00 additional. Setup time 6 AM. Auction, large flea market, refreshments available. ITT Technical Institute will hold an open house during the Swapfest. All interested are invited to tour the facilities. For additional information: Dave Johnston, K9HDQ, c/o ITT Technical Institute, 9511 Angola Court, Indianapolis, IN 46368. (317) 875-8640.

1986 "BLOSSOMLAND BLAST" Sunday, October 5, 1986. Write "BLAST", PO Box 175, St. Joseph, MI 49085.

CALIFORNIA: FCC exams, Novice-Extra. Sunnyvale VEC ARC (408) 255-9000 24 hour. 73, Gordon, W6NLG, VEC.

MASSACHUSETTS: The MIT UHF Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday, August 20, 7 PM, MIT Room 1-134, 77 Mass Avenue, Cambridge MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-0160/646-1641 or Craige Rodgers (617) 494-1986. Exam fee $4.25. Bring copy of current license (if any), two forms of picture ID and completed form 610 available from FCC in Boston (223-6609).

OPERATING EVENTS

"Things to do . . ."

OHIO STATE FAIR Special Event, August 1-17. Listen for W8TO, 80-10 meters. Exchange QTH and RST. All correspondence and requests for awards to ARS W8TO, att: State Fair event coordinator, 280 East Broad St, Columbus, Ohio 43215.

The Penn-Jersey ARC will operate W2SJT, from 1600 UTC, August 9 to 1600 UTC August 10, to honor the Oxford Furnace, an historic landmark. 2m, 450 and all HF bands. For certificate send QSL and 3 stamps to Ron Semonche, WB2TOJ, 263 W. Carlton Avenue, Washington, NJ 07882.

SCHAUMBURG SEPTEMBERFEST The Schaumburg ARC will operate club station WB9TXO from the grounds of the Schaumburg Septemberfest, August 31, 1600Z-2100Z. For certificate send QSL to SARC, PO Box 94251, Schaumburg, IL 60194.

IBM RADIO CLUB will operate special event station WD0GNK, August 16, to celebrate the 30th anniversary of the IBM Corp in Rochester, Minnesota. Certificate for QSL and SASE via WD0GNK-IBM Radio Club, IBM Corp, Dept. 868, Hwy 52 North, Rochester, MN 55901.

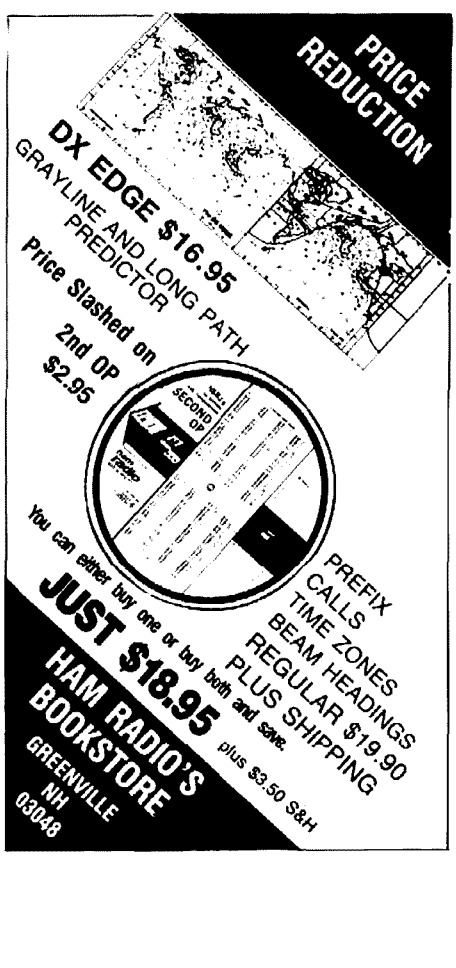
The NORTH OTTAWA ARC will operate KA8USK aboard the US Coast Guard cutter Mackinaw as part of the Grand Haven Coast Guard Festival July 30-August 3. For commemorative certificate send QSL and 9x12 SASE via NOARC, Box 44, Fennsburg, MI 49404.

NEW JERSEY QSO PARTY All Amateurs worldwide are invited to participate. 2000 UTC August 16 to 0700 UTC August 17, 1300 UTC August 17 to 0200 UTC August 18. Phone/CW same contest. Contact a station once each band. Phone/CW separate bands. Call "CQ New Jersey" or "CQ NJ". New Jersey stations sign "DE NJ". Exchange QSO number, RST and QTH. NJ stations send county. Send logs and comments to Englewood ARA, PO Box 528, Englewood, NJ 07631-0528 with 10 SASE for results.

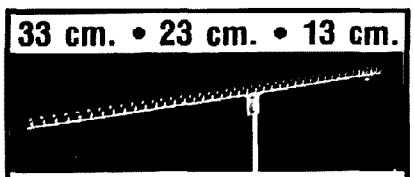
THE WILD BUNCH 160 meter SSB Contest, 0000 GMT August 2 to 2400 GMT August 3. Deadline for logs 9/7/86 postmark. Send to Rob Koziomkowski, KA1SR, 5 Watson Drive, Portsmouth, RI 02871.

DX CERTIFICATE. Nigerian Amateur Radio Society announces its "NARS at 25 Award". To encourage more contact with Nigerian Amateur Radio stations. DX stations need to work only 5 Nigerian ham stations to obtain this DX certificate. Send proof of contact plus $5.00 to NARS at 25 Award, PO Box 2873, Lagos, Nigeria, West Africa.

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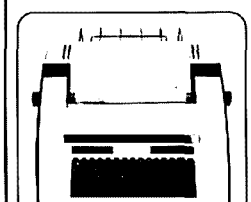


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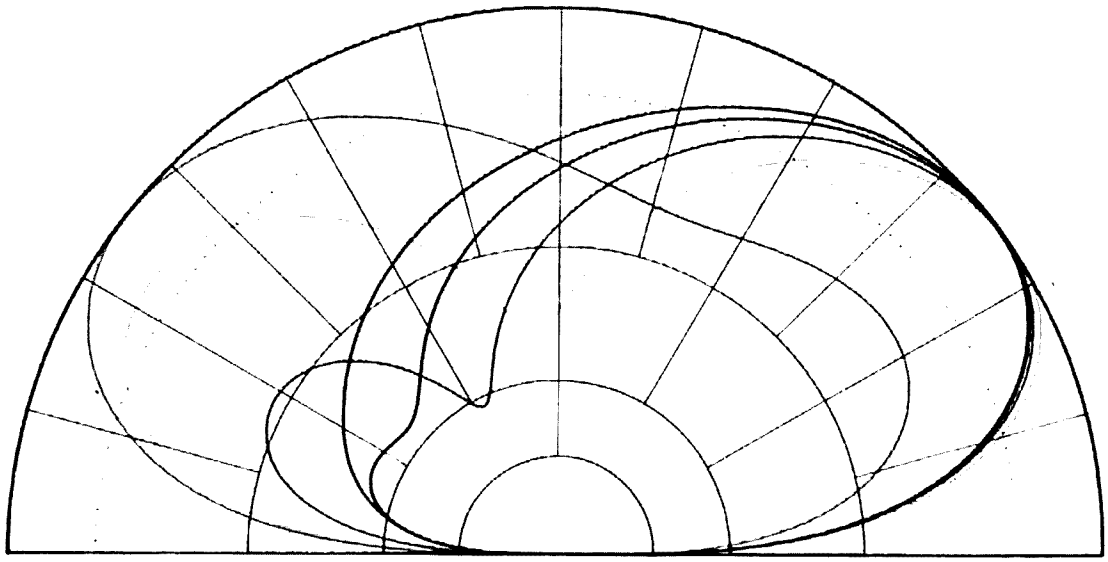
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publisher

Rich Rosen, K2RR
editor-in-chief
and associate publisher

Dorothy Rosa, KA1LBO
assistant editor

Joseph J. Schroeder, W9JUV
Alfred Wilson, W6NIF
associate editors

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editorial production

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assistant publisher

Rally Dennis, KA1JWF
director of advertising sales

Dorothy Sargent, KA1ZK
advertising production manager

Susan Shorrock
circulation manager

Therese Bourgault
circulation

cover art:
Barbara Smullen

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REFLECTIONS

switch to safety

This is an editorial that I wish never had to be written.

This past Field Day, Mike Mankey, WB0TEE, ARRL North Dakota Section Manager, was killed when the antenna he was erecting came into contact with a power line. According to the news report, the site was unfamiliar to Mike's group, and they didn't see the power lines hidden behind the trees.

I'd like to report that this is the first time something like this has ever happened, but it isn't — and I'm afraid it won't be the last, either. Perhaps as we share, with Mike's family and friends, the sorrow of his loss, it would be appropriate to give some thought to the dangers we face in pursuit of our avocation.

We've all done foolish things while working with electricity. When I was wiring 220 volts for my amplifier, for example, I was certain the circuit was dead; I'd pulled the breaker to de-energize it. But when I clamped my wire cutters and squeezed them shut, POW! Sparks, smoke, a sudden dimming of the house lights, and a wife screaming down the stairs expecting to find the worst. I was stunned and ashamed by my stupidity — I'd pulled out the wrong circuit breaker!

As I calmed both myself and my wife down, I started to analyze what I'd done wrong. It was late. I was tired and in a hurry to get the work done. I *thought* I'd taken all precautions. . . if you pull the breaker, the circuit is dead — isn't it?

I've done other very stupid things over my almost 20 years as a ham. Although I'm not proud of these experiences, I do hope I've learned something from them. WB0TEE's untimely death serves as a reminder to us all that it's imperative to always, *always* think safety, no matter where you are and no matter what you're doing.

If you're climbing a tower, make sure you're safely belted to it. If you're troubleshooting a live circuit, remove all rings, jewelry, and watches that could make you part of a complete circuit; keep one hand in your pocket. Make sure power supplies are properly built with the appropriate bleeder resistors. Don't stick your hand into your amplifier when it's turned on to "adjust" that bent plate choke. And watch out for power lines; they aren't always easy to see.

By the way, if you were to receive an incapacitating shock, would your family know what to do? Would they know not to touch you directly, but to pull you away from the electrical source with a stick or some other nonconductive tool and turn off the power ? Do they know CPR? Do they know how to get qualified emergency help as quickly as possible? Are the phone numbers posted by the phone — or better yet, on it?

We extend our deepest sympathy to Mike Mankey's family. Although we cannot lessen their loss, we can learn from it — and *switch to safety*.

Craig Clark, N1ACH
Assistant Publisher

presstop de W9JUV

OSCAR 10 IS BACK ON THE AIR ON BOTH MODES B AND L, as a result of some masterful computer analysis and reprogramming! The sophisticated bird had shut down several months ago, apparently from radiation-induced memory damage. After extensive earlier efforts to regain control failed, it seemed almost certain OSCAR 10 would be off the air for good. However, analysis of the limited responses that could be induced by control station commands provided clues to what portions of its memory still seemed intact, and when DJ4ZC uploaded software written for those memory areas the satellite came back to life!

It Now Appears OSCAR 10 Will Be Usable In Both Modes for the foreseeable future, but with some limitations; at this time the most important is a 100-watt ERP limit on uplink signals. The CW/RTTY beacon is apparently gone for good, but attitude control has been restored so the control stations should be able to keep the batteries well charged. There's cautious hope that OSCAR 10 can be kept up and running at least until the Phase 3C satellite goes up, though recent problems with the French Ariane launch vehicle have pushed its launch well into 1987. Check the various AMSAT nets for further updates.

The Japanese JAS-1 Satellite's Launch, Scheduled For Late July, was also pushed back but at presstime was still expected during the first half of August. In addition the launch of two new Russian Amateur satellites, RS-9 and RS-10, is likely in September.

AMSAT Has A New Illustrated Catalog of Membership Supplies, software, and publications. It's available from the AMSAT office (301 589-6062) or at AMSAT hamfest booths.

MIAMISBURG, OHIO'S TOXIC TANK CAR FIRE THAT BEGAN JULY 8 SPARKED one of the nation's largest mass evacuations ever and once more proved conclusively Amateur Radio's value in emergency situations. Almost 400 area Amateurs from about 20 clubs took part during the four-day emergency, providing the essential interconnecting link for participating local, state, and federal agencies. Since most agencies are limited to their own unique frequencies, and those in many cases were jammed with the volume of communications, Amateurs rode police cars, fire trucks, and even a helicopter to provide those working the disaster with the only common communications network that could tie them all together.

In Addition, Amateurs Also Set Up Weather Tracking Nets that provided firefighters with precise rain shower and wind shift information, vital to plotting the toxic cloud's path. Amateurs also coordinated the evacuation of more than 300 nursing home residents to a safe location, monitored hospital emergency room loading, and arranged food shipments to the various evacuation centers. In the aftermath of the 100-hour-long operation the Amateur communicators received very high marks from those they were assisting and as well as plenty of very positive media exposure. Congratulations to all for a very fine job!

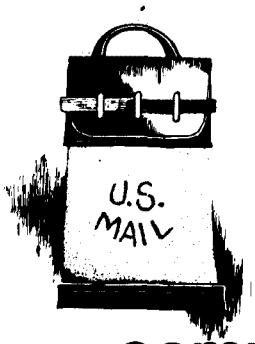
ARRL HAS ASKED THE FCC FOR IMMEDIATE ACCESS TO THE 18-MHZ WARC hand, based on belief that the U.S. government is no longer using that band. Unfortunately, the League's information wasn't accurate, so don't expect any favorable action on 18 MHz soon.

NEW EIA OPPOSITION TO THE COMMUNICATIONS PRIVACY ACT HAS RAISED HOPES that the bill may not make it through the Senate in its present form after all! In a well-thought out and strongly worded letter sent to all U.S. Senators, the Mobile Communications Divison of the Electronic Industries Association cited a number of reasons for opposing S-1667. For example, with respect to radio communications, the letter says "...it is not reasonable to attempt to legislate privacy..." so that the bill would, in effect, "...instill a false expectation of privacy!" The letter then points out that the bill is inconsistent with the Communications Act of 1934, which recognizes that a prohibition against intercepting radio communications is impractical but establishes severe sanctions against the misuse of intercepted information. It further notes the bill is in itself inconsistent, since it admits that cordless telephone transmissions are easily intercepted -- so can't be protected -- but provides unqualified protection to cellular transmissions! "This discrimination," the letter continues, "certainly cannot be rationalized on a technical basis....There is no justification for this legislation to treat one mode of radio communications differently from others!" Another point: "There can be no true expectation of privacy with radio communications unless the message is encrypted (scrambled)." It concludes, "The nature of radio communications cannot be changed by legislation..." but concedes that the problems cited could be changed by amendment.

Support For The Latter Is Far From Unanimous Within The EIA, and at least one land mobile manufacturer has sent key senators a followup letter stating its opposition to the EIA position. Nevertheless, the EIA makes valid points, and the letter's arrival appears to have seriously compromised the bill's hopes of the bill's supporters for quick Senate passage.

Amateurs And Others Who Oppose S-1667 In Its Present Form should call or write their Senators to register support for the EIA Mobile Communications Division letter.

"CAPTAIN MIDNIGHT," WHO BROKE IN ON AN HBO MOVIE IN APRIL to object to scrambling of satellite TV, has been caught. He's John Mac Dougall, a 25-year-old satellite TV dealer and a part-time operator for an Ocala, Florida uplink station. Unfortunately, he's also KA4WJA, so some news reporters have also been condemning Amateur Radio along with Mac Dougall! In a plea-bargaining agreement he's been fined $5000 and sentenced to one year's probation with his Technician Class license suspended for a year.



comments

happy endings

Dear HR:

I've been meaning to write for some time now to let you know of the response to my letter published in the October, 1985, issue of *ham radio*. I started a letter several times, but didn't know exactly what I wanted to say. A simple "thank you" wasn't quite enough.

I received many nice cards and letters and have made many new friends. The greatest surprise came late one evening when I received a call from a representative of ICOM America who reported seeing my letter and asked whether I would be interested in an ICOM transceiver. Then she asked the question that left me speechless: if I had a choice of any ICOM radio, which would I choose? "A 751," I replied. When she told me one would be shipped the following day, I honestly didn't believe it. I thought it was some kind of joke.

When the UPS truck arrived with two large packages just two days later, I realized it was no joke, nor was it a dream. I opened the larger box and found a brand-new ICOM 751 complete with a factory-installed PS35 power supply, FL52A filter, voice synthesizer, hand mic, and all the power cables. *WOW!* In the other box was an ICOM 251A, 2-meter, all-mode transceiver with all the accessories and an SM8 Duo Desk microphone. *Double WOW!*

Is this the end of the story? No, far from it. Many hams responded with other equipment and good wishes. I have tried and believe I have responded to each person who contacted me, but I'd still like to express my sincere and deep appreciation to everyone who assisted me in any way, whether it was with equipment or just a kind word of encouragement.

And of course, my sincerest, heartfelt thanks go to those caring people at ICOM America.

My dream, from the beginning, has been to provide phone patches for missionaries in foreign countries. I'm on the air now and able to provide assistance in handling written traffic, but two more items are needed: a phone patch and an amplifier. God has truly blessed me in enabling me to become part of the fraternity of Amateur Radio. I hope that by providing phone patches I'll be able to give someone a small token of pleasure or service they might not otherwise have had.

At 32, I don't know what a toll MS will take on me, but I can assure you it will have a good fight. Somehow I feel that the day I give up or give into the limitations — if there really are any — will be the day my dream ends. That day must not come; I must not let it.

In the two years since I became disabled I've realized that I, like many, once found too many excuses for accepting limitations. But I've also disovered that a limitation is only an obstacle between you and your goal. If you can't go through it, then you'll just have to go around, under, or over it. The point is to reach your goal, any way you can ...

John Statham, N5HTQ
McComb, Mississippi 39648

no problem!

Dear HR:

How often have you purchased an item that later developed a problem — and then found out that getting the item fixed was an even *bigger* problem?

I bought a Mirage amplifier and was pleased with its performance. But one day, after a particularly long-winded QSO, the power amp in my B3016 2-meter amplifier quit in a puff of smoke, despite its good heat sink and an attached blower.

I sent the unit, which was a couple of years old, back to Mirage for repair. Needless to say, I was surprised to have it returned promptly, with "NO CHARGE" marked on the invoice. I had expected a long exchange of purchase receipts, letters, charge card information, and the like. What a surprise to have a warranty repair done quickly and without a quibble!

I use three Mirage amplifiers, two at home and one mobile, and have found them all to be well made and long-lasting, even in the 100 percent duty cycle of ATV operation. I would recommend Mirage to anyone who wants a quality product.

Henry B. Ruh, KB9FO
Chicago, Illinois

narrowband filters

Dear HR:

Two versions of the program described in my March, 1986, article, "Build Narrowband RF Filters," are now available for computers other than those originally mentioned.

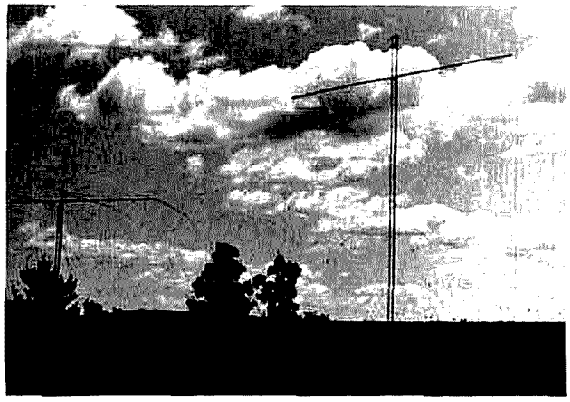
Atari owners may contact Marion D. Kitchens, K4GOK, 2709 Colt Run Road, Oakton, Virginia 22124. Marion will provide a copy (on disk). Owners of the Timex 1000 can contact Rudy Knaack, W7FGQ, 11415-28th S.W., Seattle, Washington 98146. Rudy will provide a listing for an SASE, or will copy the program on cassette tape for $5.

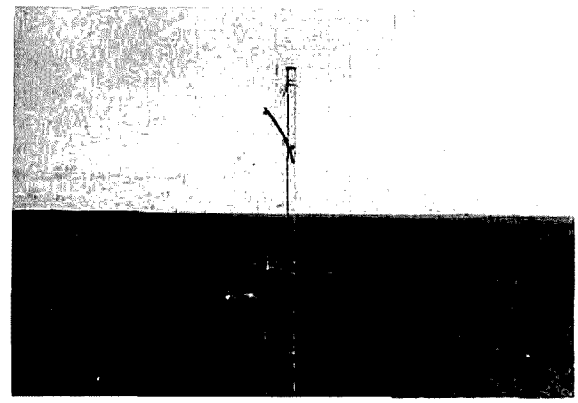
I might add that both versions improve on my method of calculating Chebychev order and save program lines.

The ultimate line-saving version of the program comes from Ken Stringham, AE1X, who uses the equations that derive k and q values that I entered in table form (**lines 1000-1500**). Although Ken didn't mention that he'd provide listings or copies, he did send me a copy. Someone interested in adapting the program to another machine might be interested in getting this listing (written in C-64 BASIC) from me.

I have had inquiries from owners of the IBM PC (and clones), but as yet have no information for them. If anyone is working on translating my program to any other machine, I'd appreciate knowing. I'll act as a "clearing house" for information about the program and sources.

Bob Lombardi, WB4EHS
Melbourne, Florida 32935





Ground and aerial views of two element 80m delta loop array supported by 46′ boom at 115′.

analyzing 80-meter delta loop arrays

NEC program assesses performance in the presence of "real" ground

During the early part of 1985, I finally managed to complete the installation of a "gain" antenna for the 80-meter band: a reflector-driver parasitic array with two equilateral triangular loop elements. Later in the year, I realized another ambition: the ability to analyze HF wire antennas over real ground with a trustworthy computer program. In this article, which describes the results of applying that capability to the analysis of my new antenna system, I'll show the behavior, over poor ground, of the peak gain and "average" front-to-back ratio across the band, along with a number of radiation patterns. I'll describe the antenna system itself in just enough detail to define the item under study.

Of course, the first question one always asks about any project is "How can I make it better?" Toward this end, I'll show, at a single frequency, the performance obtainable with a number of more-or-less feasible variations on the existing design, including the following:

- alternative feedpoints;
- higher boom;
- closer element spacing;
- both elements driven; and
- square loops.

I'll also discuss a few other "gain" parasitic antenna systems, including pairs of dipoles and inverted vees, and the half-wave sloper. Finally, I'll show the effect of a less convenient option — relocating to the seacoast — that provides the best results of all.

As suggested above, this isn't a construction article. Neither is it intended to present a design procedure. Even with the power available in current microcomputers, the number of trials required to find an "optimum" design is prohibitive unless one narrows the scope at the outset. This is the purpose of this article: to provide data on which to base an informed selection of basic design choices.

the antenna

After years of careful deliberation (i.e., procrastination), I ultimately decided to build a system very similar to the W2PV array that I discussed in a previous article.[1] The two identical equilateral loops are supported, apex up, by a 46-foot boom at the 115-foot level on a 120-foot guyed tower. The loop's circumference is 264 feet (88 feet per side). I used No. 10 AWG copper wire, primarily because it was already "in stock."

Figure 1, a diagram of the configuration, shows the coordinate system that I'll use in defining radiation patterns. The *x* axis is parallel to the boom from the reflec-

By Bill Myers, K1GQ, Box 501, Hollis, New Hampshire 03049.

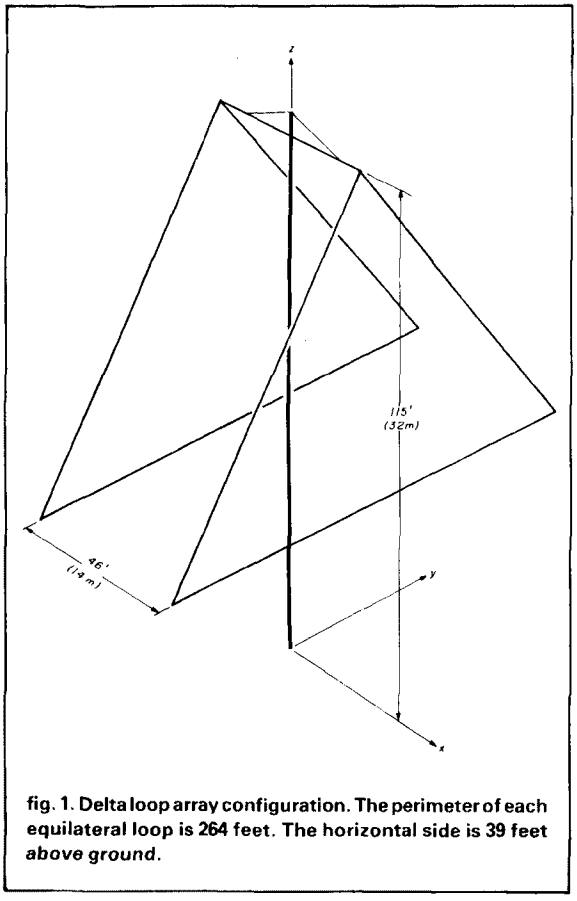


fig. 1. Delta loop array configuration. The perimeter of each equilateral loop is 264 feet. The horizontal side is 39 feet above ground.

tor end toward the driven element. The *y* axis lies in the ground plane and in the plane of the loops, and the *z* axis is positive upward.

The two loops are configured as a reflector plus driver parasitic array by adding a 4 μH coil in series with the reflector element. An additional 12 μH is connected in series with each loop to switch the center frequency from 3.8 to 3.5 MHz. The necessary switching is accomplished using relays in a small box at each loop apex.

Equal lengths of RG-11A/U connect these boxes to a third relay box at the center of the boom. I use chokes (patterned after the W2DU balun[2]) at these feedpoints to reduce currents on the outside of the feedlines. The central switchbox selects either or both loops to be connected to the main transmission line, which is 75-ohm CATV cable.

On 80 meters, transmission line losses are very small, even when the line is operated at high SWR (i.e., mismatched at the antenna). For simplicity, I decided not to attempt any matching at the antenna. Instead I use lumped-reactance networks at the station end, with additional relays to select different networks so that I can instantly switch between matched parasitic or driven configurations.

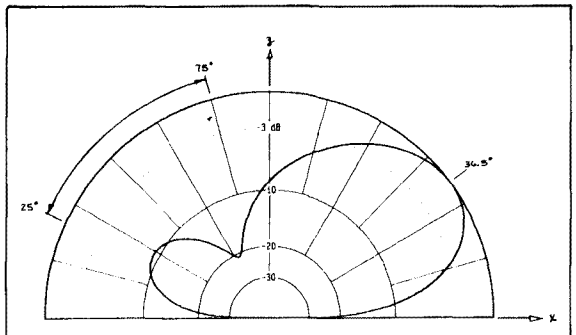


fig. 2. Elevation radiation pattern at 3.8 MHz. Peak gain is 8.69 dBi at 36.5 degrees. Front to back averaged over 25 to 75 degrees (see text) is 11.49 dB. The x and z axes are illustrated in fig. 1.

the computer program

I'm fortunate to have access to the resources needed to run the complete Numerical Electromagnetics Code (NEC).[3] This program is ideally suited to analyzing the performance of wire antennas up to a few wavelengths in size, with arbitrary shapes and excitations, in the presence of "real" ground. For those familiar with the method of moments, I modeled the six straight wires that make up the two-element array with 11 segments each. Excitation was applied at the centers of the two segments surrounding the apex, since the NEC models sources at the segment centers rather than at the segment junctions. The ground characteristics were dielectric constant, 4, and conductivity, 0.001 S/m, the usual values for poor ground such as the rocky, hilly terrain behind my house.

radiation patterns

The NEC program produces an enormous listing of numbers that's difficult to fully assimilate. Fortunately, I've succeeded in compressing much of the information, without too much agony, into a more visual format — namely, polar radiation pattern plots. It's important to understand these plots well, so I'll explain an example of each of the two types of pattern charts in some detail.

Figure 2 shows the chart I'll refer to as the *elevation pattern*. The thick curve on the inside of the grid is the ratio (in dB) of the total radiation intensity, in the direction defined by the angular coordinate of the chart, to the radiation intensity that would be obtained if the antenna input power were uniformly distributed over a sphere. Most folks simply call this ratio "gain relative to isotropic," with the notation "dBi" when the ratio is expressed in decibels. The angular coordinate for this chart is measured in the vertical plane containing the boom; that is, the *xz* plane shown in **fig. 1**. The grid interval on the chart is 15 degrees, although the resolution for pattern data points is *one* degree.

The outermost radial of the elevation pattern grid corresponds to zero dB, relative to the peak gain *in the xz* plane. The value of peak gain is noted in each figure caption. This normalization creates good detail in the pattern characteristics. Notice also that the radial coordinate is non-uniform, as was first recommended by K1TD[4]. My charts are drawn with the scaling:

radius = $10^{G/40}$

where G is the normalized gain in dB. This scaling expands variations near zero dB so that the 3-dB beamwidth is easy to identify, without compressing variations in the sidelobe regions excessively.

The second type of radiation pattern plot is the *azimuth pattern*, as illustrated by **fig. 3**. Here the angular coordinate is measured in the horizontal plane — the *xy* plane of **fig. 1**. Although the elevation chart shows only the above-ground hemisphere, the azimuth chart must show a full circle to allow for possible asymmetries in the radiation pattern (for example, those caused by feeding the loop array at one of the lower corners).

Each azimuth pattern corresponds to a particular value of the elevation angle. The obvious choice for the elevation angle is that corresponding to the peak gain on the elevation chart. This angle is 36.5 degrees at 3.8 MHz for my delta loop array, but it varies from case to case. In order to reduce the number of computer runs, I used 36 degrees for the elevation angle in all azimuth charts presented in this article.

Most readers will have encountered the terms "H-plane" and "E-plane" in connection with radiation patterns. I'm avoiding these terms because their definition is more likely to cause confusion than illumination. Also, I've deliberately omitted labels for the angular coordinates in the radiation pattern charts because the standard coordinate system (as defined by the IEEE[5]) is counter-intuitive for those of us who grew up in an azimuth-elevation world.

peak gain

The peak gain of an antenna system is the value of the gain in the direction of its maximum value. For most of the results discussed here, the direction of the maximum occurs in the vertical plane containing the boom. Thus, unless otherwise noted, the values quoted for peak gain are to be understood as the peak in the *xz* plane.

It is not my intent, in this article, to provide an exhaustive discussion of peak gains. There are two reasons for this. First, the number of interesting cases is overwhelming when one considers the range of applicable heights, frequencies, and ground characteristics in addition to the enormous variety of antenna designs. Second, there are many well-established myths that must be systematically dealt with in any such undertaking; doing so would expand the scope of this article substantially. Thus, the gains presented herein should not be extrapolated beyond the conclusions stated below.

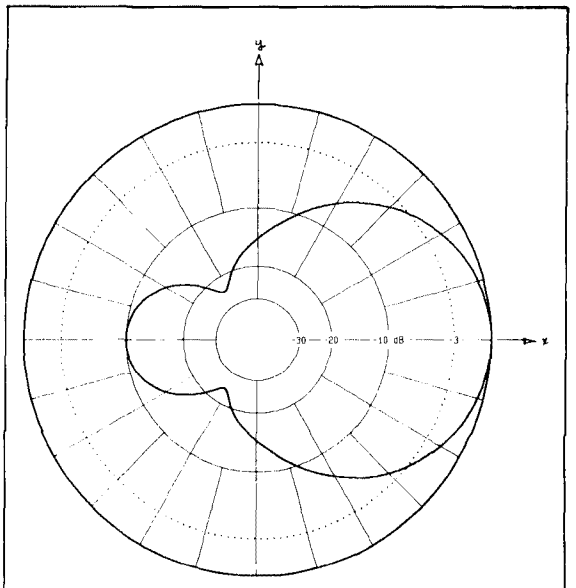


fig. 3. Azimuth radiation pattern at 3.8 MHz for 36 degree elevation angle. The x and y axes are illustrated in fig. 1.

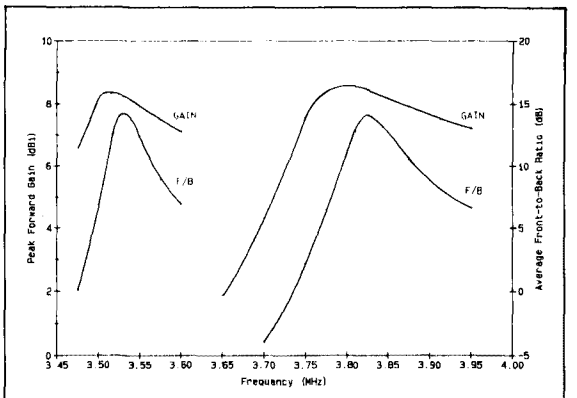


fig. 4. Two element delta loop array performance versus frequency for phone and CW modes. The phone-mode design frequency is 3.8 MHz. The CW-mode is created by series-loading each loop with a 12 μH inductor.

average front-to-back ratio

The concept of front-to-back ratio needs some elaboration to provide a useful measure of 80-meter antenna performance. As suggested by **fig. 2**, the elevation pattern can exhibit deep nulls at selected elevation angles, while showing large lobes at other angles off the back. Thus, the ratio of forward gain at a specified elevation angle to the backward gain at the same angle can vary dramatically for different elevation angles.

On the 80-meter band, *all* of the elevation angles are

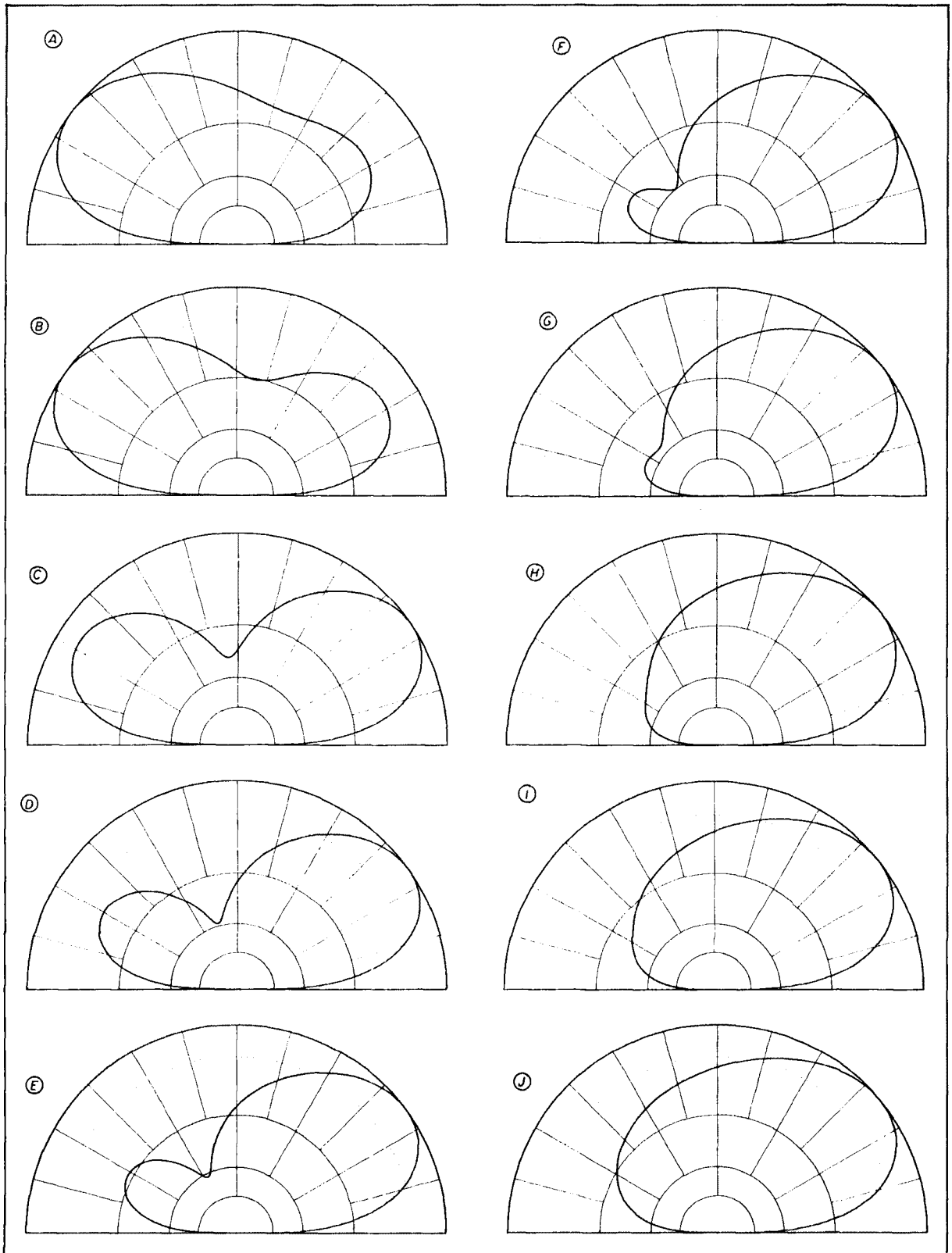


fig. 5. Phone mode elevation patterns from 3.66 (A) through 3.95 MHz (J). See table 1.

significant, from the extremely low angles needed for very long DX paths to the nearly straight-up angles required for accepting (or rejecting) local signals. As far as I can determine, the question of which angles are most important, say for the path from New England to Western Europe, remains unanswered. G6XN[6] claims that the only valid rule is "the lower the better." There seems to be ample evidence to the contrary, however. For example, many DXers in the Northeast have observed that high-angle antennas often perform better than low-angle antennas when the band first opens toward Europe in the evening. (The same observation applies to both 40 and 160 meters, as well.) In any case, 80-meter antenna systems with poor gain near the horizon (such as the delta loop array described here) are good overall performers, suggesting that angles in, perhaps, the 20- to 50-degree region are indeed useful for medium-range DXing.

In assessing the radiation pattern, we're usually interested in the ability of the antenna system to reject strong signals coming from the backward direction, which are most likely to arrive at relatively high elevation angles (the lower angles corresponding to longer paths and thus to weaker signals). Thus, I've chosen to display the ratio of the peak forward radiation intensity to the backward radiation intensity averaged over the elevation angles from 25 to 75 degrees. This power ratio is presented in the customary dB. I'll abbreviate this performance characteristic as "averaged f/b," rather than "peak forward power to backward power averaged over 25 to 75 degrees in the *xz*-plane." I believe this quantity is a more useful indicator of 80-meter antenna performance than the customary f/b ratio; furthermore, the values presented below seem to agree well with on-the-air observations.

impedance

Knowledge of the antenna input impedance is useful for two reasons. First, one can design a matching network to transform the input impedance to the characteristic impedance of the transmission line, thereby minimizing line losses due to mismatch. Second, the NEC calculation of gain requires an accurate value for the input impedance (in order to determine input power).

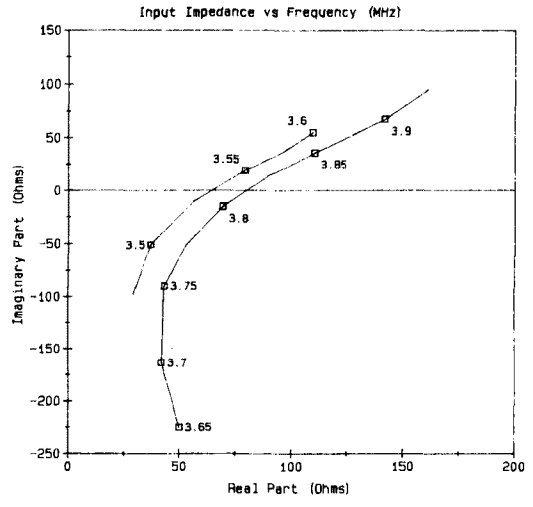


fig. 6. Computed input impedence versus frequency for phone and CW modes. These do not correlate well with measurement (see text).

The fidelity of the NEC input impedance computation is affected by the number of segments specified for the antenna model. More segments improve the impedance result (up to the onset of numerical difficulties), but also increase the run time dramatically. However, if the segment length is held constant from case to case, then the relative results can be trusted even though the absolute values may be suspect. In all of the cases presented here, the number of segments per straight wire section was selected to yield approximately the same physical segment length (8 feet).

Table 1. Phone-mode elevation patterns vs. frequency.

Frequency (MHz)	Peak Gain (dBi)	Angle (deg)	Average F/B (dB)
3.65	7.44*	38.0	−5.35
3.7	7.62*	36.0	−3.69
3.75	7.66	34.5	2.41
3.775	8.48	35.5	6.73
3.8	8.68	36.5	11.49
3.825	8.54	37.5	14.33
3.85	8.27	38.0	12.94
3.875	7.99	38.5	10.73
3.9	7.73	38.5	9.04
3.95	7.30	39.5	6.86

*In the backward direction

performance

Figure 4 shows the computed performance versus frequency for my two-element array. The phone-mode peak gain reaches its maximum (8.68 dBi) right at the design frequency, 3.8 MHz. The peak in the average f/b is about 25 kHz higher. The CW-mode gain reaches its maximum just above 3.5 MHz, and the offset to the f/b peak is also about 25 kHz. The overall response appears slightly narrower for the CW mode, as should be expected for loaded loops, but the difference in bandwidth between modes is unimportant. The maximum CW peak gain (8.40 dBi at 3.525 MHz) is slightly below the phone mode maximum.

The gain and f/b peaks can be aligned at the same frequency by at least two methods. The reflector loading

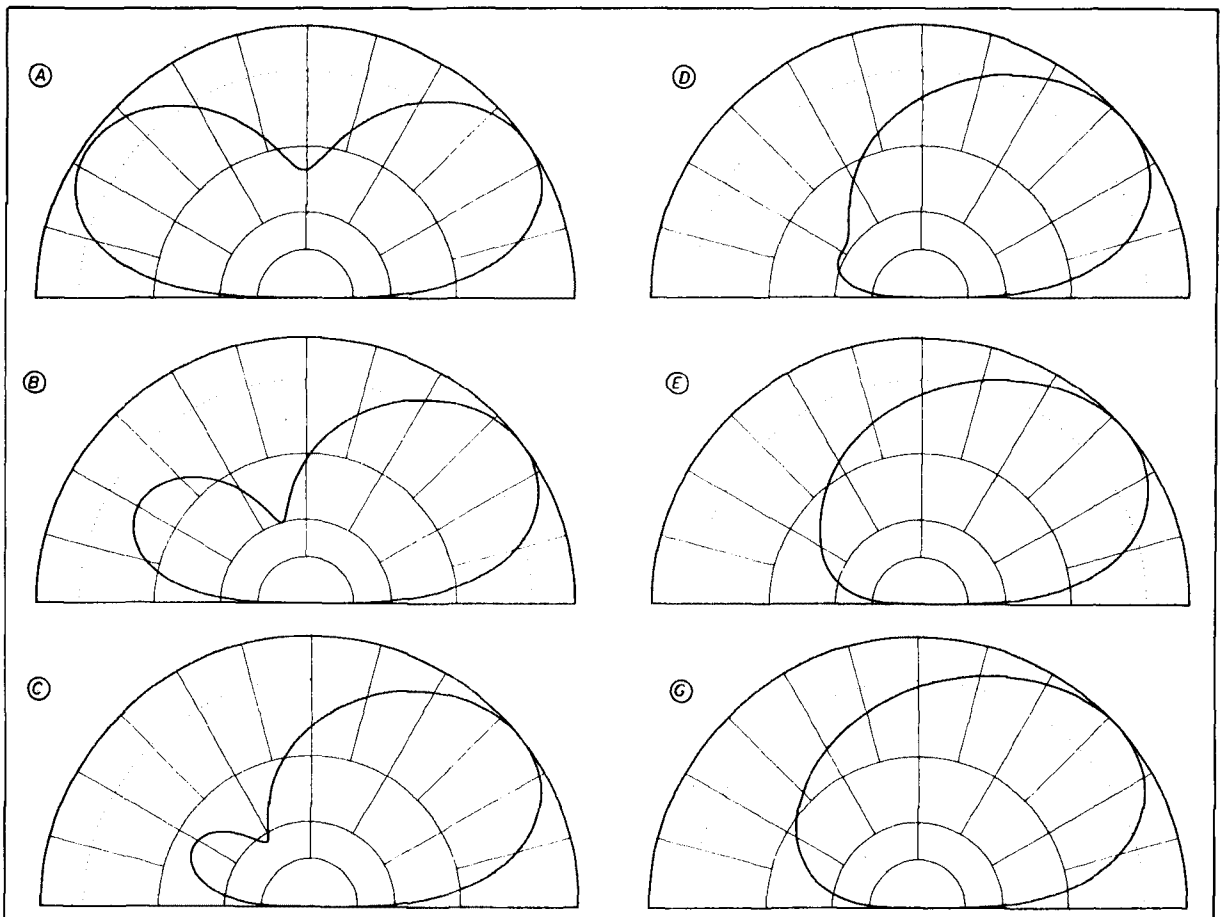


fig. 7. CW-mode elevation patterns from 3.475 (A) through 3.6 (G) MHz. See table 2.

inductance can be varied. Or one can simply redefine the range of angles over which the average is computed. The point here is that the 25-kHz offset is a silly amount to worry about.

A more significant result is the pattern reversal just below 3.75 MHz. Negative f/b ratio means that the direction of the peak gain is toward the reflector. This occurs because the parasitic element is electrically too short to act like a reflector. The potential for disaster caused by too-short loops is obvious; if I ever rebuild the loops I will probably increase the circumference 4 feet to move the design frequency down about 50 kHz.

Figure 5, a series of elevation radiation pattern plots at a number of frequencies surrounding the design frequency, shows the pattern reversal quite nicely (see also **table 1**). Note that the elevation angle of the main lobe is nearly the same at every frequency, increasing uniformly from 34.5 degrees at 3.7 MHz to 38.5 degrees at 3.9 MHz. If we define the *effective height* of an antenna system as the height of a horizontal half-wave dipole, above perfect ground, which yields the same elevation for the first lobe of the ground reflection pattern, then:

Table 2. CW-mode elevation patterns vs. frequency.

Frequency (MHz)	Peak Gain (dBi)	Angle (deg)	Average F/B (dB)
3.475	6.67	34.5	0.38
3.5	8.28	36.5	7.14
3.525	8.40	39.0	14.31
3.55	8.01	40.5	12.42
3.595	7.57	41.5	9.16
3.6	7.21	42.0	7.21

$$h_e = \lambda/(4 \sin\theta)$$

where θ is the elevation angle. The effective height for the parasitic array (at 3.8 MHz) is 0.42λ, or 109 feet. Thus, even though most of the antenna structure is well below the effective height, the array still yields a relatively low angle for the main lobe.

This rather surprising result comes about because of the supression of higher angle radiation by the array free-

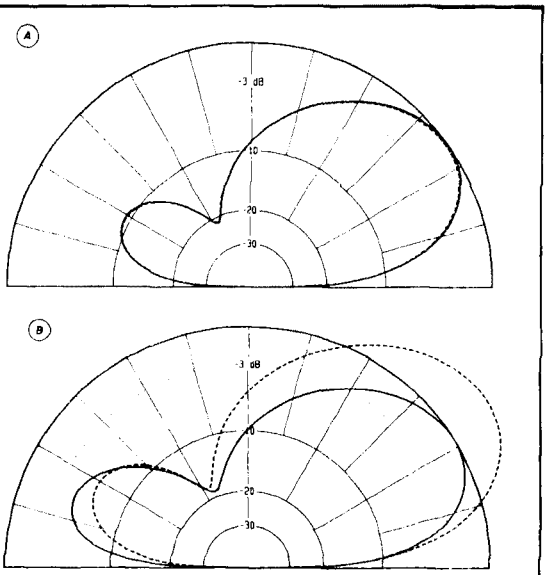


fig. 8. Elevation patterns at 3.8 MHz for alternate feedpoints. Peak gains are 8.80 dBi and 5.48 dBi for (A) bottom and (B) corner feed, respectively. Corresponding average F/B ratios are 11.53 dB and 7.83 dB. The dashed pattern on these charts is for top feed.

space directivity. If the directions for the peak gains of the free-space pattern and the ground reflection pattern were aligned, we could expect gain in the neighborhood of 13 dBi (7 dBi from the array and 6 dB from the reflection). The actual peak gain is lower partly because the lobes are not aligned (the free-space array peak is on the horizon, where the ground reflection pattern has a null) and partly because the reflection isn't perfect.

The input impedances calculated by NEC for the delta loop array are plotted in **fig. 6.** The phone mode system is resonant near 3.815 MHz, with the resistive component equal to about 80 ohms. The CW mode resonance is about 3.535 MHz, at 65 ohms. These results don't correlate with SWR measurements or with admittance measurements at the transmission line input. This isn't too surprising, however, because the antenna model I provided as input to NEC isn't especially realistic in details that can influence impedance, such as adjacent towers and uneven terrain.

Figure 7 shows the CW mode elevation patterns for frequencies at the bottom of the band (see also **table 2**). The elevation angles for the gain peaks are slightly higher than for phone mode, but not enough to be of concern. The best pattern occurs a bit high in the band for an inveterate DXer, but I've found on-the-air performance to be excellent at the very bottom of the band.

In fact, in practice the apparent f/b ratio has been generally consistent with that indicated by the performance calculations. On both modes, S5 stations in Europe disappear into the band noise when the array direction is switched. Louder Europeans decrease in signal strength about 2 to 3 S-units, while US stations come up by the same amount. My receiver is a TS-930; I've checked the S-meter calibration and found it to be about 5 dB per S-unit. The array shows essentially no f/b at 3.75 MHz and is definitely backward (in the phone mode) at 3.7 MHz.

I currently have two other 80-meter antennas, a dipole at 120 feet parallel to the array boom, and a full-size quarter-wave vertical with 12 radials. Neither of these antennas should work well in the directions favored by the delta loop array . . . and they don't. On the other hand, the dipole is better to the Caribbean, South America, and Japan. While the loops do exhibit some rejection "off the side," it's not as much as the dipole exhibits off its ends. Overall, I'm quite pleased with the agreement between the theoretical results and actual performance, especially considering that the terrain surrounding my antenna system is far from planar.

bottom and corner feed

My loops are fed at the top. Full-wave loops can be fed anywhere on the circumference with no appreciable change in the input impedance or in the *free space* directivity. However, the selection of the feedpoint has a potentially devastating effect over real ground. Two interesting alternative feedpoints are at the center of the horizontal bottom side and at either of the two corners. The elevation patterns for these two cases, at 3.8 MHz, are shown in **fig. 8**. The bottom-fed pattern is indistinguishable from the top-fed pattern (the dashed curve) and the peak gains are nearly the same: 8.80 dBi versus 8.68 dBi, respectively. (This small difference may well

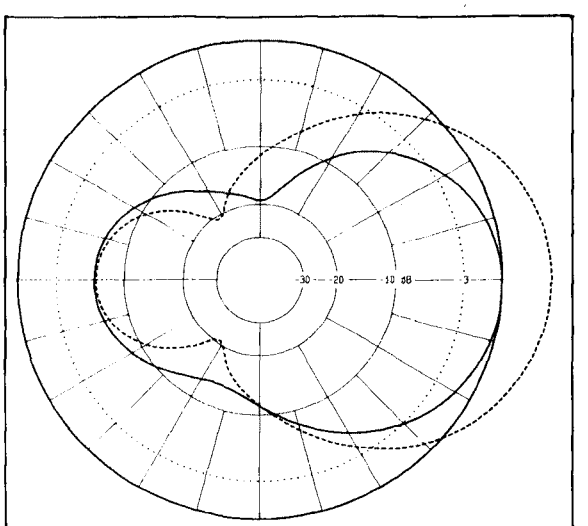


fig. 9. Azimuth pattern at 3.8 MHz with corner feed. The pattern is skewed towards the side of the boom containing the feedpoint. The dashed pattern is for top feed.

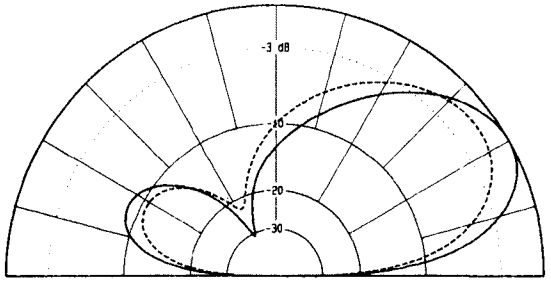


fig. 10. Elevation pattern at 3.8 MHz with the array raised to 145 feet. Peak gain is 9.97 dBi at 30.5 degrees and average F/B is 11.10 dB. The dashed pattern is for boom at 115 feet.

be due to modeling errors that lead to small input impedance errors and thus to peak gain errors.) The corner-fed pattern, on the other hand, is clearly inferior in the backward direction; furthermore, the peak gain is dramatically reduced to 5.48 dBi. The corner-fed azimuth pattern (**fig. 9**) shows that the pattern is skewed and the side null is filled in on the side of the boom corresponding to the driven corner.

The explanation for the inferior performance of the corner-fed arrangement is simple. The amplitude of the reflection coefficient for poor ground is much lower for vertical polarization than for horizontal polarization. Thus, the "gain" produced by ground reflection is smaller for vertically-polarized radiated fields. The corner feed produces a substantial vertically-polarized component in the total field, whereas the top and bottom feeds produce entirely horizontal polarization, so the corner feed has less gain (see Appendix for additional discussion).

increased height

Another feasible modification to my delta loop array would be to raise the antenna. Since I guy my towers every 30 feet, the next "natural" boom height above 115 feet is 145 feet. **Figures 10** and **11** show the elevation and azimuth patterns at 3.8 MHz for this configuration. Interestingly, while the depth of the rearward null is increased, the magnitude of the rearward lobe is also increased and the average f/b is almost unchanged. The peak gain is quite a bit higher, up 1.29 dB to 9.97 dBi, and the elevation angle at the peak is 6 degrees lower, at 30.5 degrees. The change in the angle is not too important, but the potential for increased gain, which is rather hard to come by, may be worth the challenge of constructing and maintaining a 150-foot tower.

closer spacing

I chose an unusually wide spacing for my array because I guessed that the increase in radiation resistance

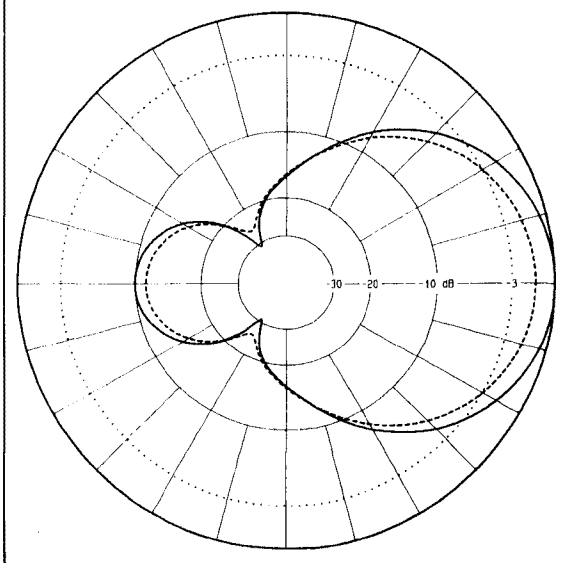


fig. 11. Azimuth pattern at 3.8 MHz with the array raised to 145 feet. The dashed pattern is for boom at 115 feet.

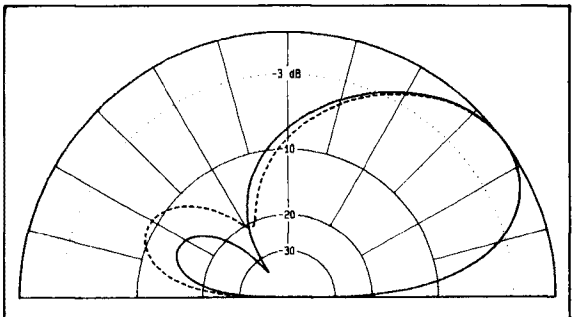


fig. 12. Elevation pattern at 3.8 MHz with spacing reduced to 32.4 feet. Peak gain is 8.70 dB at 36.5 degrees and average F/B is 16.07 dB. The dashed pattern is for 46-foot spacing.

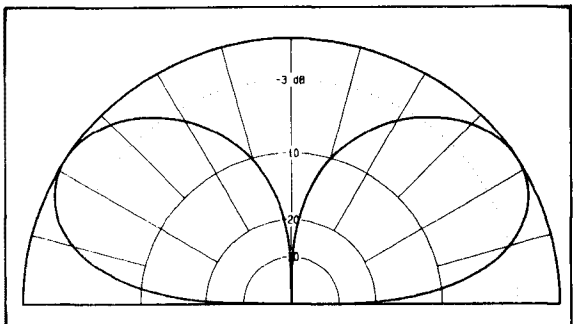


fig. 13. Elevation pattern at 3.8 MHz with both loops driven 180 degrees out-of-phase. Peak gain is 7.05 dB at 33 degrees.

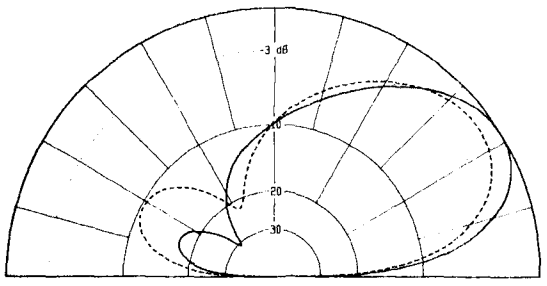


fig. 14. Elevation pattern at 3.8 MHz for square loops with the same perimeter as the triangular loops. Peak gain is 9.71 dBi at 31.5 degrees and average F/B is 16.52 dB. The dashed pattern is for triangular loops.

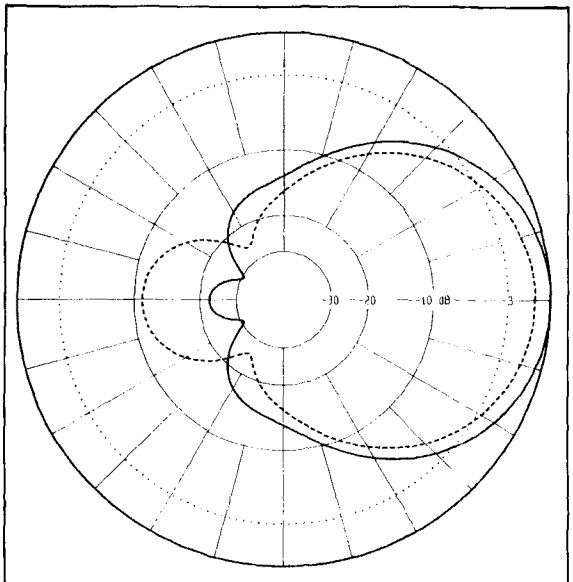


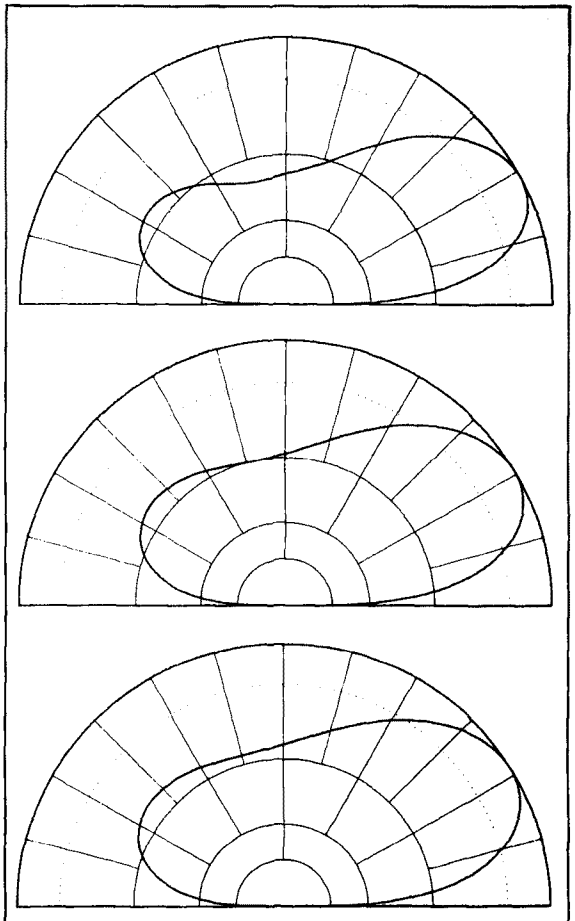
fig. 15. Azimuth pattern at 3.8 MHz for square loops. The dashed pattern is for triangular loops.

with increased spacing would partially offset the relatively lower value created by proximity to ground. Also, I expected that the bandwidth would be somewhat better at the larger spacing. Because the array is parasitic rather than driven, performance isn't terribly sensitive to spacing. Although I haven't verified all of these conjectures, I did examine performance for spacing reduced from 46 feet to 32.4 feet at λ/8 (3.8 MHz). The elevation pattern shown in **fig. 12** has nearly the same peak gain (8.70 versus 8.68 dBi) at the same elevation angle, but the backward pattern is significantly better. The input resistance drops 19 ohms while the input reactance increases 15 ohms.

both loops driven

As mentioned earlier, I designed the remotely controlled switches so that I could drive both loops together if I chose to. The feedpoint boxes are identical, and installed facing each other, so paralleling the feedlines at the center of the boom results in exactly out-of-phase drives to the two loops (with the feedlines from boom center to loop apex being equal in length).

A two-element array driven with 180-degree phase shift produces an end-fire pattern that's independent of spacing over a reasonable range near λ/8. The elevation pattern has a very deep null straight up, as seen in **fig. 13**. I thought this null might produce good effective f/b when listening to Europeans, assuming that signals from



Element Type	Peak Gain (dBi)	Angle (deg)	Average F/B (dB)
dipole	9.26	28.5	6.92
120° vee	8.54	31.0	6.62
90° vee	8.04	32.0	6.47

fig. 16. Elevation patterns at 3.8 MHz for halfwave elements.

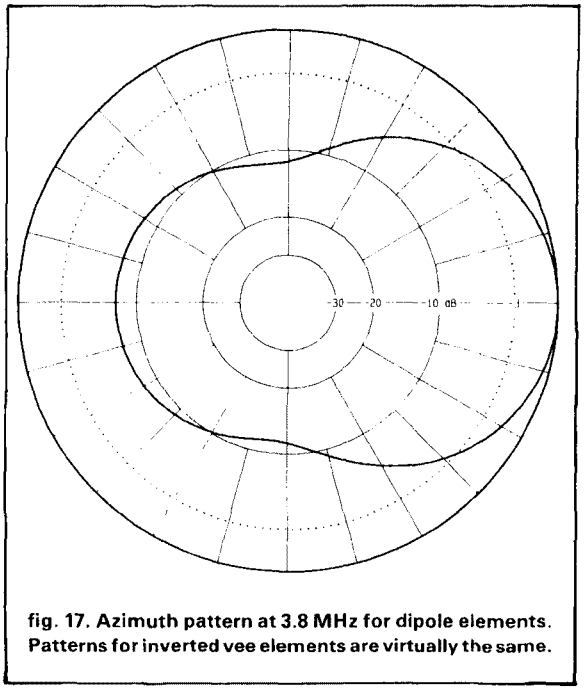


fig. 17. Azimuth pattern at 3.8 MHz for dipole elements. Patterns for inverted vee elements are virtually the same.

close-in W's would arrive at high elevation angles. However, in practice I found that the pattern of the parasitic array (**fig. 2**) rejected local signals better than the driven array.

The calculated peak gain for the driven array is 7.05 dBi. This is significantly below the gain of the parasitic array. Furthermore, the input resistance is much lower, so that the effect of losses in the CW loading coils becomes more significant (however, these coils aren't required for a driven array). I found that signals from Europe were never better with the driven arrangement, and occasionally were noticeably worse (note that 1.5 dB is generally *not* measurable except with laboratory instrumentation). I've abandoned this setup.

square loops

The mechanical challenges inherent in constructing an array of square loops in place of delta loops are severe but not outrageous. According to G6XN[6], we can't expect that the change in the shape of the elements will yield any appreciable change in gain, but the pattern may be improved (i.e., the side lobes can be decreased) with increased mutual coupling. Changing to square loops with the top horizontal side at the same height as the apex of the delta loops raises the effective height, which will increase peak gain and decrease the elevation angle at the gain peak.

These conclusions are supported by the results. The peak gain for the elevation pattern shown in **fig. 14** is 9.71 dBi, about 1 dB higher than the delta loop array. The elevation angle at the peak is 5 degrees lower, at 31.5

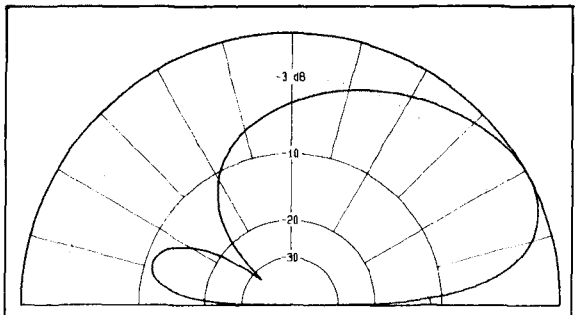


fig. 18. Elevation pattern at 3.8 MHz for a halfwave sloper plus 120 foot grounded tower. Peak gain is 1.56 dBi at 28.5 degrees and average F/B is 11.12 dB. Peak gain without the tower is 0.81 dBi at 31.5 degrees.

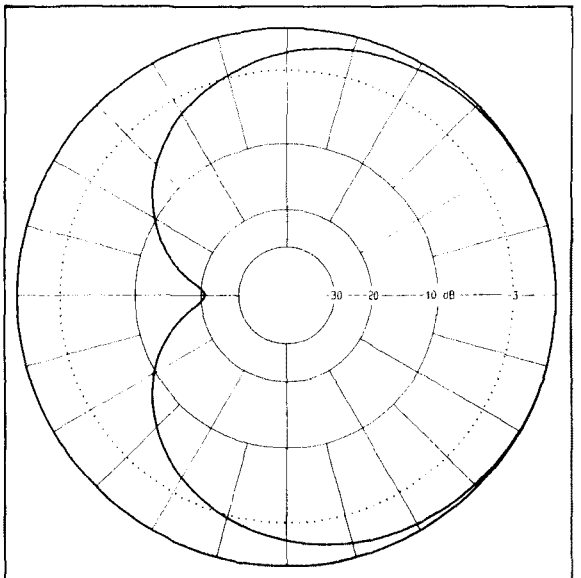


fig. 19. Azimuth pattern at 3.8 MHz for halfwave sloper plus 120 foot grounded tower.

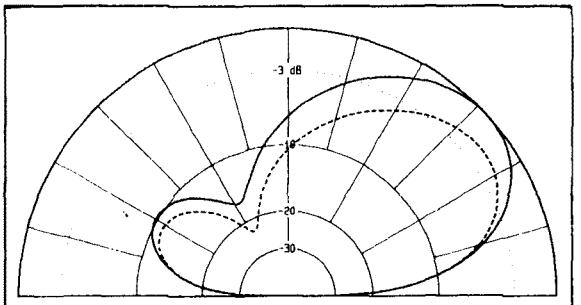
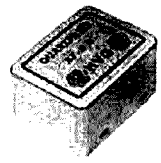


fig. 20. Elevation pattern at 3.8 MHz for sea water ground conditions. Peak gain is 10.47 dBi at 42.0 degrees and average F/B is 11.25 dB. The dashed pattern is for poor ground.

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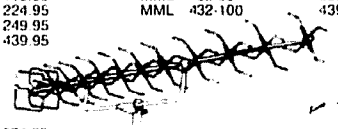
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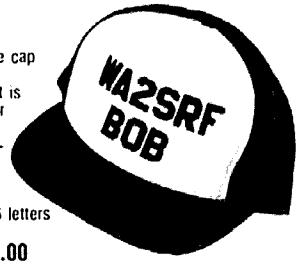
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degrees. And the pattern is outstanding; the average f/b for this case is 16.5 dB! The azimuth pattern is shown in **fig. 15**; this pattern is also excellent.

half-wave elements

Erecting closed one-wavelength loops on a guyed tower can be a major hassle when there are trees within shouting distance of the tower. (By the way, I have ample experimental evidence that shouting at trees in hope of convincing them to release captured wires is ineffective.) Because half-wave elements are much simpler to handle, I examined the performance of three two-element parasitic arrays with:

- horizontal half-wave elements at 115 feet;
- inverted-vee half-wave elements from 115 feet, with 120-degree apex angle; and
- inverted vee elements with 90-degree apex angle.

Figure 16 shows the elevation pattern charts for these three cases. There's a tendency for the pattern null to fill in as the apex angle decreases. However, even the horizontal dipole array pattern is inferior to the delta loop pattern (**fig. 2**). It may be possible to "tune up" the pattern by changing the element spacing and the loading inductance for the reflector, but I haven't attempted this analysis. The azimuth pattern for the horizontal dipole case is given in **fig. 17**; the patterns for the two inverted vee cases are almost the same.

On the other hand, the peak gains are competitive with the delta loop array: 9.26, 8.54, and 8.04 dBi, respectively, versus 8.68 dBi for the loops. It appears that the inverted vee array with wires as flat as possible is a good substitute for the delta loops as far as gain is concerned.

the half-wave sloper

My previous 80-meter antenna system was a set of four half-wave slopers, slanting about 30 degrees from the vertical. Unfortunately, I had to dismantle this system before the delta loop array was operational, so I have no on-the-air comparison of the delta loops versus the slopers.

Figure 18 illustrates the elevation pattern for a single half-wave sloper (slanted 30 degrees from vertical), in the presence of a 120-foot tower grounded at its base. The backward rejection is rather good — the average f/b is 11.1 dB. The corresponding azimuth pattern is included in **fig. 19**, even though it's unremarkable.

Now for the bad news: the peak gain (at 28.5 degrees) is a paltry 1.56 dBi. Compare this with a simple inverted vee from 120 feet with 120-degree apex angle, which exhibits a peak gain of 6.18 dBi at 31 degrees. Once again, the source of this inferior performance is polarization. The field in the plane containing the sloper and the tower is entirely vertically polarized, which is disastrous over poor ground (see Appendix). In fact, the peak gain of the sloper over poor ground is *less* than the same antenna

in free space (2.16 dBi). One possible explanation might be that the proximity to ground distorts the nominally-sinusoidal current distribution on the sloper so badly that the directivity is degraded. However, examination of the NEC printout for this case shows that the current is symmetrical about the feedpoint (the center) within a few percent in amplitude and within 1 degree in phase. Thus, the degradation must be due to destructive interference between the (attenuated) reflected field and the direct field.

Clearly, the sloper — like the corner-fed loop — is a poor choice for a transmitting antenna at sites with poor ground characteristics. I'm planning to reinstall one of my slopers to verify whether or not this computed 7-dB disadvantage appears in practice.

performance over good ground

I've disparaged the performance of vertically-polarized antennas over poor ground, but the fact is that poor ground degrades the performance of *any* type of antenna for 80 meters at other than grazing angles. **Figures 20** and **21** show the radiation patterns for my delta loop array when moved to an island surrounded by salt water. The pattern nulls are filled in somewhat and the elevation angle for peak gain is *increased* to 42 degrees. However, the peak gain is considerably increased, by nearly 2 dB, to 10.47 dBi. This figure probably represents the maximum gain achievable with a two-element array of delta loops at 115 feet.

summary

I've presented polar plots, gains, and average f/b data for a number of 80-meter antennas. **Table 1** collects all of the results in one place and includes the input impedances computed by NEC. These latter data are not reliable because I did not follow the procedures needed to confirm that enough segments were provided to assure that this calculation had "converged" to the true value. However, all cases were run with comparable segment lengths, so the trends in impedance should be representative of the true behavior.

The significant column in **table 3** is the peak gain; it's possible that the average f/b for any of the variations could be tweaked somewhat by modifyiing the reflector's resonant frequency and the element spacing.

The primary conclusions of this study of 80-meter antennas over poor ground are:

- The delta loop array provides enough gain to be worth the effort, along with truly useful f/b ratios, even though the boom height is too low by conventional wisdom.
- Raising the boom 30 feet would add about 1.3 dB of gain at a slightly lower elevation angle.
- Changing from delta loops to square loops would increase gain about 1.0 dB and would improve the pattern.

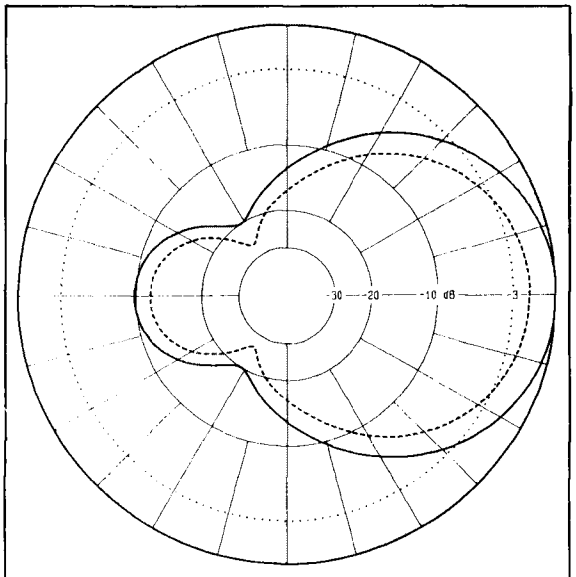


fig. 21. Azimuth pattern at 3.8 MHz and 36 degree elevation angle for salt water ground. The dashed pattern is for poor ground.

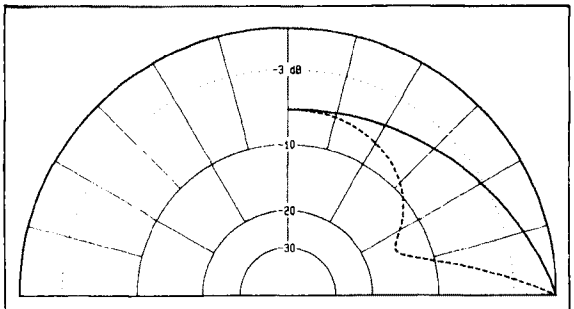


fig. 22. Reflected field intensity amplitude versus angle of incidence for horizontal (solid curve) and vertical (dashed curve) polarization with poor ground ($\epsilon_\rho = 4$ and $\delta = 0.001$ S/m).

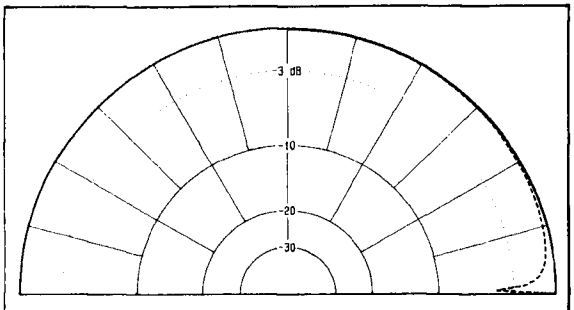
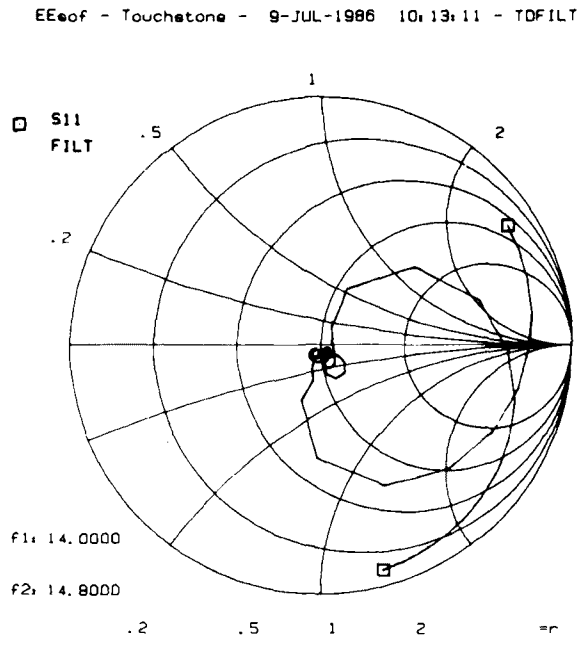


fig. 23. Reflected field intensity amplitude versus angle of incidence for horizontal (solid curve) and vertical (dashed curve) polarization with good ground ($\epsilon_\rho = 80$ and $\delta = 5$ S/m). The solid curve nearly coincides with the 0 dB arc at all angles.

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Table 3. Summary of Antenna Cases.

Description	Peak Gain (dBi)	Elevation Angle (deg)	Average f/b (dB)	Input Impedence (ohms)
GQ Delta Loop Array	8.68	36.5	11.5	69.2 j15.3
GQ at 3.5 MHz	8.28	36.5	7.1	37.0 -j51.3
GQ bottom fed	8.80	36.0	11.5	129.2 -j23.6
GQ corner fed	5.48	33.0	7.8	71.4 -j57.3
GQ 30' higher	9.97	30.5	11.1	80.1 j12.0
GQ on 32.4' boom	8.70	36.5	16.1	50.2 -j0.0
GQ over salt water	10.47	42.0	11.3	55.5 -j11.5
GQ with square loops	9.71	31.5	16.5	119.5 +j41.0
GQ with both driven	7.05	33.0	0.0	25.2 -j44.3
Two dipoles	9.26	28.5	6.9	78.2 +j20.1
120 deg Inverted vees	8.54	31.0	6.6	64.6 +j5.2
90 deg Inverted vees	8.04	32.0	6.5	45.8 -j21.3
Sloper & tower	1.56	28.5	11.1	103.1 +j22.3

- Moving the feedpoint to the bottom center would have essentially no effect; moving it to a corner would be disastrous.
- Shortening the boom to λ/8 would lower the radiation resistance and improve the pattern somewhat, but would not change the gain.
- Replacing the loops with dipoles or flat inverted vees would not affect gain very much, but the pattern would be degraded.
- My half-wave sloper system should not have worked.
- I should retire to an island off the coast of Maine.

acknowledgements

Many thanks are due John Kenny, W1RR, and Doug Grant, K1DG, who constructively reviewed this article.

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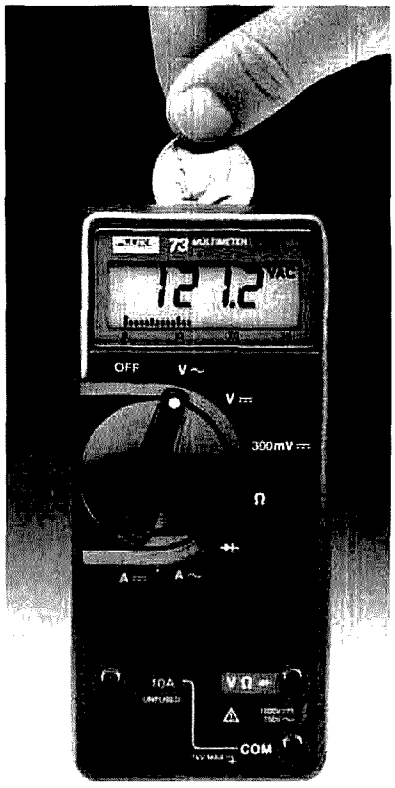
appendix
ground reflection amplitudes

The effects of real ground on Amateur antenna installations have been reported many times, so I don't propose to reiterate those analyses. However, I did develop an unusual polar plot which helps in clarifying some of those effects. To understand the chart, a bit of prefatory explanation is needed.

The presence of ground is modeled by constructing a *reflected field* that is vectorially summed with the direct field of the antenna. The reflected field from a given point on the ground is proportional to the field incident from the antenna; the proportionality constant is called the reflection coefficient. The reflection coefficient depends on polarization of the incident wave and on the angle of incidence, as well as the characteristics of the ground (which depend on frequency). This rather complicated modeling problem is simplified somewhat by decomposing the incident field into horizontal (parallel to the ground) and vertical (in a plane normal to the ground) components and applying separate horizontal and vertical reflection coefficients. Expressions for these coefficients are given in many texts; for example, Chapter 4 of Ma's *Theory and Application of Antenna Arrays*.[7]

The two reflection coefficients are complex numbers; that is, the reflected field components are modified in both amplitude and phase. The effect of phase shifts is a modification of the positions of peaks and nulls in the final pattern, which is developed by constructive and destructive recombinations of the incident and reflected fields. The effect of amplitude modification is to change the

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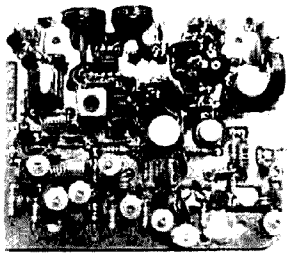
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levels of the peaks and nulls. For example, if the reflected wave amplitude is one-half of the incident wave, the combined field will have peaks no greater than 1.5 times the amplitude of the original field (+3.52 dB). The maximum combined field with perfect reflection is 2 times the original amplitude, or 6.02 dB, so 2.50 dB of "potential" gain has been lost.

The horizontal and vertical reflection coefficients are rather messy functions of angle of incidence, dielectric constant, conductivity, and frequency. To illustrate the behavior of these parameters at 3.8 MHz, I've plotted the reflection amplitude versus angle of incidence in the same format I used for the elevation patterns. That is, the radial coordinate represents the amplitude of the reflected radiation intensity (power) relative to the incident radiation intensity, and the angular coordinate corresponds to the elevation angle for the incident field. **Figure 22** shows this chart for poor ground, with dielectric constant = 4 and conductivity = 0.001 S/m. The solid curve is the amplitude for the horizontal reflection coefficient and the dashed curve is for the vertical reflection coefficient. Note that both are unity (zero dB) for grazing angle incidence, and both are equal for normal incidence. The very pronounced null in the vertical reflection coefficient amplitude shows graphically why vertical polarization is inferior to horizontal polarization over poor ground, except at extremely low angles. (To be fair, it must be noted that at low antenna heights, vertical polarization is superior — even though the amplitude of the vertical coefficient is always less than that of the horizontal component. This is because the phase shift for the horizontal polarization guarantees destructive recombination, whereas the recombination for vertical polarization is constructive in the region above the null.)

This situation changes dramatically over good ground. **Figure 23** shows the same chart for salt water, with dielectric constant equal to 80 and conductivity equal to 5 S/m. Both coefficients are essentially unity for all angles of incidence, except for the narrow null near grazing incidence for the vertical coefficient. The advantage of horizontal over vertical polarization does not occur in this environment.

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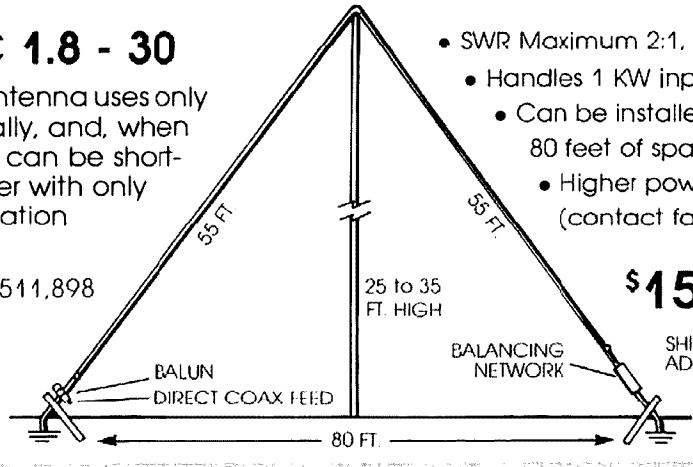
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This brief article describes the construction of a relatively high-performance RF amplifier that uses a Signetics NE5205 wideband RF amplifier. The NE5205[1] is an integrated wideband RF amplifier that serves as an excellent general-purpose RF gain block for applications from a few Hertz to above 600 MHz. It provides a non-inverting 20-dB gain and, although not an LNA, its typical 6-dB noise figure (50-ohm input) is quite good and should be adequate for many RF projects.

The NE5205 is available in either a TO-46 metal package or an eight-pin small-outline (SO) package. Although samples were distributed in the conventional eight-pin mini DIP ("N" package), that package isn't shown in the NE5205 literature and doesn't appear to be available from distributor stock.

The SO package is a very inexpensive plastic unit with the same operating characteristics as the TO-46. A minor problem with the SO package, however, is size; it's about one-twentieth the volume of the more common eight-pin mini DIP. The leads, on 0.05-inch centers and bent at the ends in the form of small feet, are much too short to go through even a thin PC board. This is a surface-mount device (SMD) designed to be mounted on the trace side of the PC board or other substrate. The package is small, but it isn't difficult to handle; you will, however, have to be quite careful and you may have to buy a small tip for your soldering iron.

NE5205 features

The functional schematic for the NE5205, shown in **fig. 1**, is considerably more complex than typical integrated wideband RF amplifier gain blocks. The small die size, offering low propagation delays and low parasitic elements, in part give this component its good operational characteristics. A particularly useful feature is the use of several different feedback loops to stabilize the gain and operating point and provide good input and output impedance matching. While this is too lengthy a topic to be covered here, the NE5205 data sheet provides complete details.[1]

Don't try to build the circuit shown in **fig. 1** from discrete components; such a construction might work to a few tens of megahertz, but in general its performance will be very poor at best. The success of this circuit design depends as much on the small die size and integrated construction as on the actual circuit configuration. In any event, it's much less expensive to use the NE5205 than to build a discrete version.

In a minimum basic circuit, the NE5205 needs only three external components: input and output coupling capacitors and a supply bypass capacitor. Adding a few inexpensive components, however, results in a more versatile amplifier with much better tolerance for the abuse of experimenting than a minimum parts-count amplifier would offer.

By Michael E. Gruchalla, 4816 Palo Duro NE, Albuquerque, New Mexico 87110

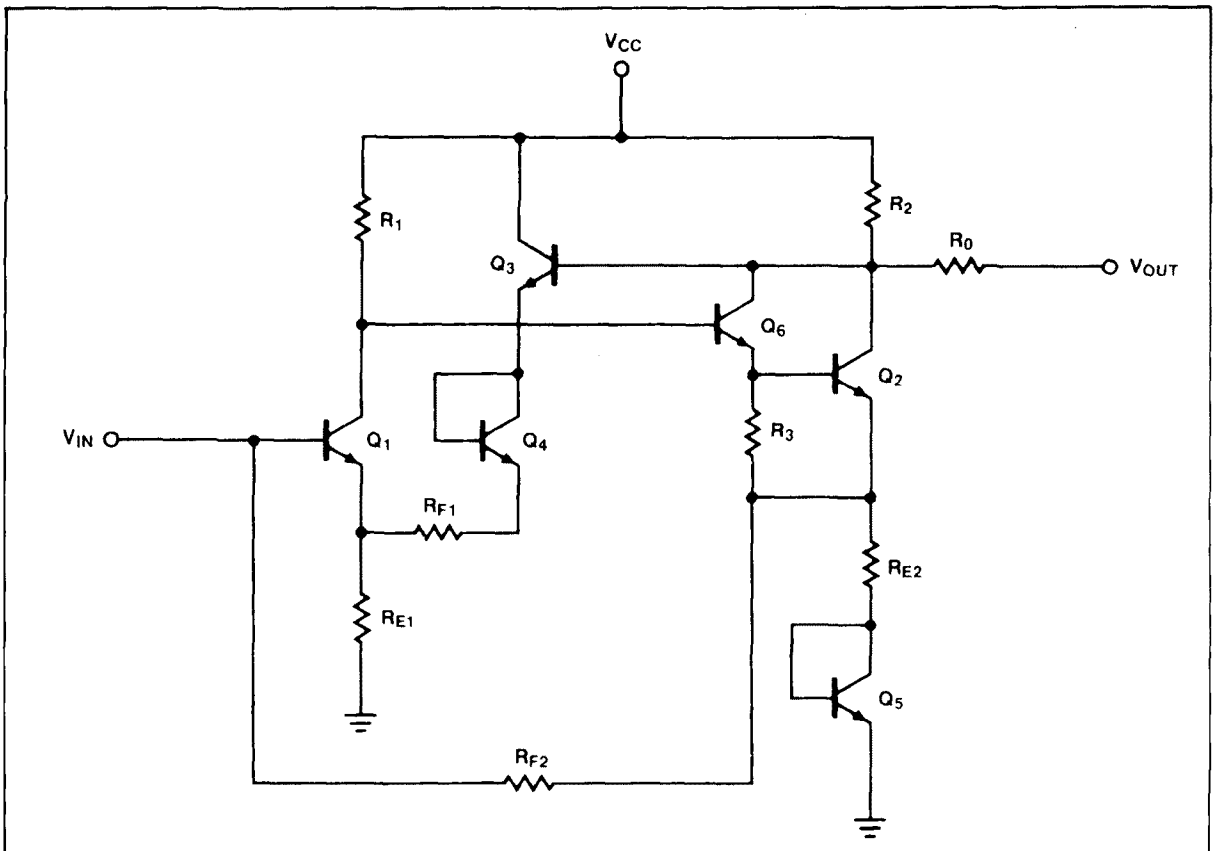


fig. 1. NE5205 functional diagram.[1]
(Copyright 1985, Signetics Corporation used with permission.)

amplifier circuit

Figure 2 shows the schematic of the complete amplifier. The input diode, CR1, prevents reverse-bias damage to the input transistor of the NE5205. The input can generally tolerate short pulses of high forward bias since the input transistor is simply forward-biased. However, even short pulses of reverse (negative) input bias as low as a few volts will avalanche the input emitter-base junction, totally destroying the part or seriously degrading its noise performance. When CR1 is included, it will conduct with a reverse input drive of about 1.4 volts, limiting the input level to the NE5205 to a safe value. This diode is reverse-biased at about 1 volt by the input potential of the NE5205. Since the maximum usable input of the NE5205 is about 80 mV peak (− 12 dBm), the 1-volt reverse bias of CR1 prevents its conduction with normal signals and helps reduce the diode capacitance to minimize loading of the input. CR1 is a Schottky diode with a faster response speed than the NE5205 and it has minimum capacitance. If you can't find the 1N5711 device specified for CR1, a 1N4148 or 1N914 will work reasonably well; the high-frequency cutoff may be reduced, however. CR1 may simply be omitted, but the finished amplifier will then be quite sensitive to input damage.

Resistors R1, R2, and R3 form a Pi pad to allow a little adjustment of the amplifier gain and input impedance. They also provide some additional input protection. Similarly, R4, R5, and R6 form an output pad for gain and output impedance adjustment. These two pads may be omitted for maximum gain, but I find that keeping about a 0.5-dB pad at each place provides better stability in general applications. The resistors should be a high-quality carbon composition type. Low-noise carbon film or metal film may be used if their high-frequency performance is adequate.

The coupling capacitors, C1 and C2, must be dipped tantalums to minimize the physical size and provide good high-frequency performance. Don't try to use aluminum electrolytics; if you do, you're likely to end up with an oscillator. Capacitors C3 and C4 are supply bypass capacitors. High-frequency bypassing is provided by C3, with C4 providing low-frequency

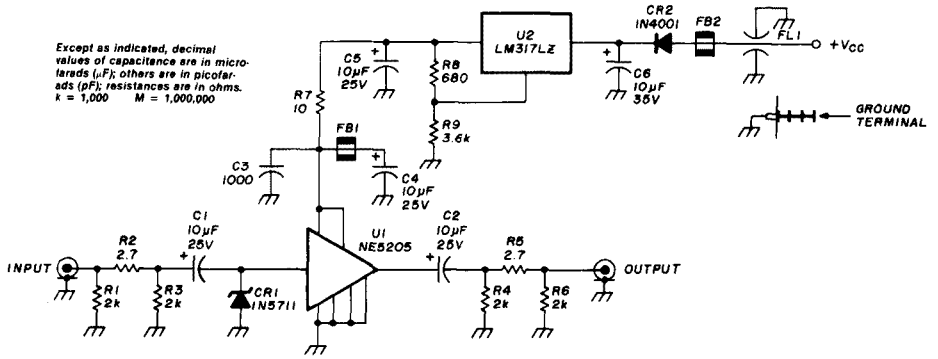


fig. 2. Amplifier schematic diagram.

bypassing. The ferrite bead, FB1, prevents the lead inductance of C4 from forming a parallel resonance with C3. Here the bead is acting like a very lossy inductance that totally destroys the Q of the C3/C4 resonant circuit. Without the bead, the parallel resonant network formed by C3 and the leads of C4 will cause the effective power supply impedance to be high at the resonant frequency. This can cause an artifact in the frequency response and in some cases can cause instability (oscillation) due to coupling into the amplifier's internal circuitry through the power supply paths.

Resistor R7 helps to decouple high frequencies from the power supply pin, improving stability with cascaded amplifiers. With its associated components, regulator U2 protects the amplifier from over-voltage damage and stabilizes operation with unregulated supplies. It too may be omitted if a good 5- to 8-volt regulated supply is available, but it's generally more convenient to provide regulation as part of the basic amplifier design.

Diode CR2 protects the amplifer from the application of a reversed power supply voltage, probably one of the more common types of damage to experimental circuits. The power input filter FL1 prevents noise from coupling into the amplifier from the power supply leads. Bead FB2 on the filter lead provides a little additional filtering of the power supply input.

choosing the correct PC board material

The 3:1 artwork for the amplifier is shown in **fig. 3**. This layout is tailored for installation in the enclosure discussed later; if you choose to use a different enclosure, you'll have to make some alterations to the mounting details, but don't change the basic circuitry unless you're experienced in RF PC board layout.

This board was designed for surface-mounting of all components even though only the SO-package NE5205 is a true surface-mount part. The PC board should be constructed of 1/32-inch, 2-ounce double-clad epoxy-glass material. Standard PC board doesn't make a good substrate for surface mounting because the pads will pull off easily, particularly during soldering. However, if you're careful, you can achieve quite adequate results. The only component holes necessary are those that tie to the ground plane. Since there are no component holes other than those grounded, there's no need to clear the ground plane from any of the component leads on the ground-plane side of the board. Only single-sided PC board construction is then required. A full ground plane is needed on the nontrace side, but no etching is required on the ground plane. This makes the board a bit easier to duplicate and consequently a little less expensive if you have it fabricated.

Though it's easier to have the PC board fabricated, it's actually simple enough to make using a cut-and-peel technique. A small hobby knife can be used to

Box	*Pomona 3751 or similar (Newark Electronics No. 34F1260/$18.50)*
C1,C2	*Capacitor, 10 μF/25V dipped tantalum —*
C4,C5	*Sprague 196D or similar*
C3	*Capacitor, 1000 pF CK05 or disc*
C6	*Capacitor, 10 μF/35V dipped tantalum — Sprague 196D or similar*
CR1	*Schottky diode, 1N5711 (may substitute 1N914 or 1N4148 if the 1N5711 cannot ba obtained)*
CR2	*Rectifier diode, 1N4001 or equivalent*
FB1,2	*Ferrite bead (Amidon FB-43-101 or equivalent)*
FL1	*Filter, Erie 1250-003 or equivalent (Newark Electronics No. 10F8145/$3.13)*
PCB	*PC Board, per* fig. 3
R1,R3,R4, R6	*Resistor, 2 k, RC07 carbon composition — Allen Bradley or equivalent*
R2,R5	*Resistor, 2.7 ohm, RC07 carbon composition — Allen Bredley or equivalent*
R7	*Resistor, 10 ohms, RC07 carbon composition — Allen Bradley or equivalent*
R8	*Resistor, 680 ohms, RC07*
R9	*Resistor, 3.6 kilohms, RC07*
U2	*Regulator, LM317LZ*
Terminal	*Ground Terminal, (Newark Electronics No. 40F6026/$15.00/100 each)*
U1	*RF amplifier, Signetics Electronics, NE5205D*

Note: Jameco Electronics carries many of the above items and can supply most of the components (except the PC board) on an "After Receipt of Order" basis subject to their line-item and invoice minimums and distributor availability.

A complete kit for this project is available from Radiokit, Box 411H, Greenville, New Hampshire 03048. Contact Radiokit for details.

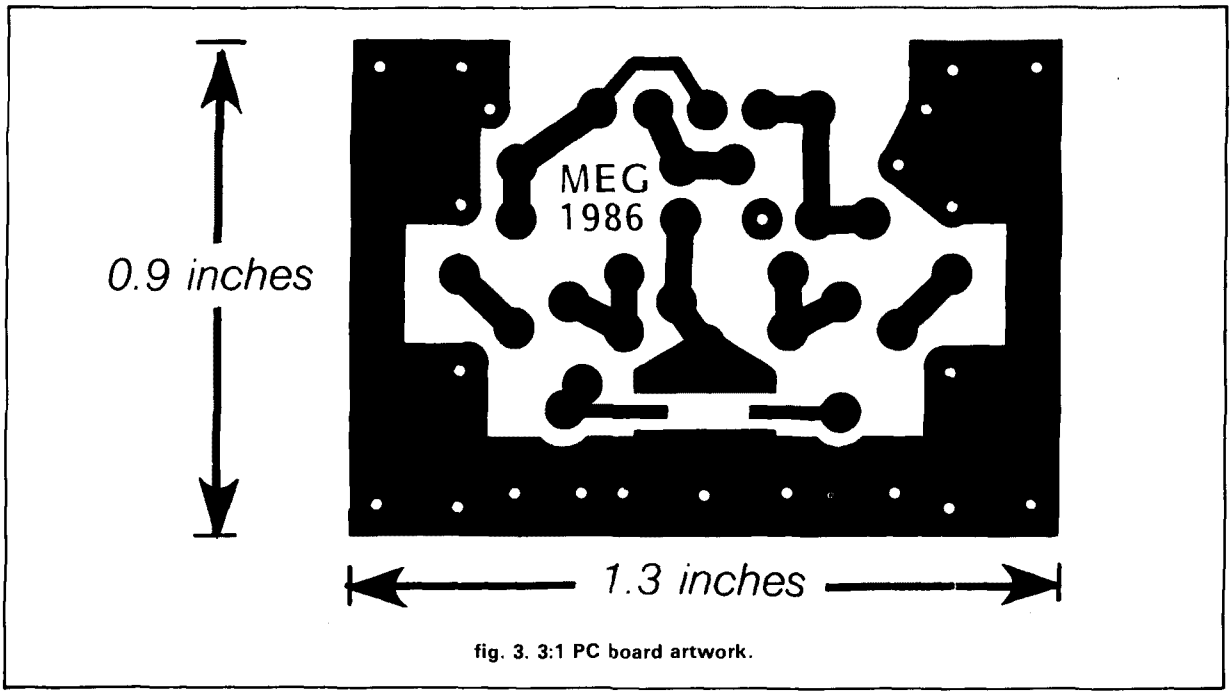


fig. 3. 3:1 PC board artwork.

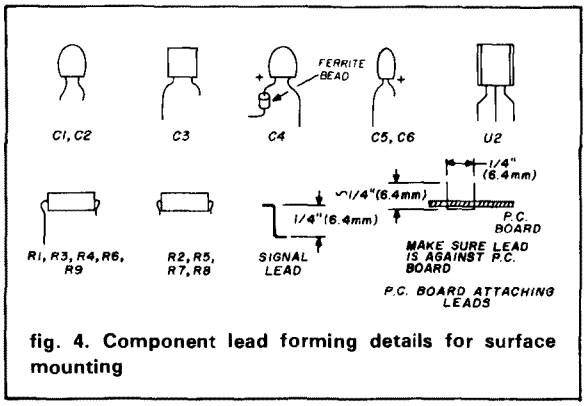


fig. 4. Component lead forming details for surface mounting

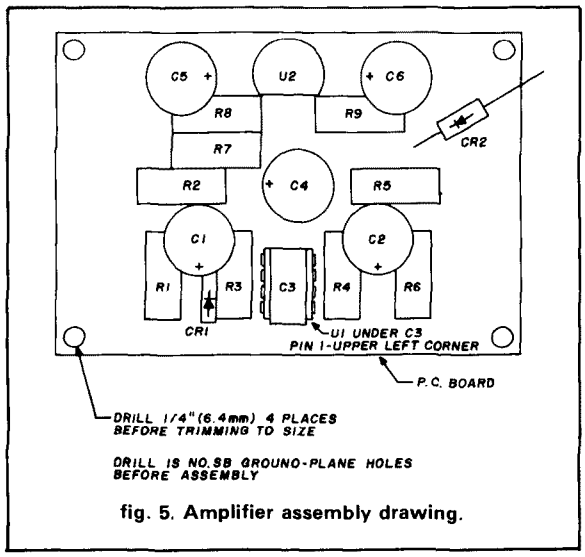


fig. 5. Amplifier assembly drawing.

cut the trace outline through the cladding on one side of a precut piece of double-clad PC board and a needle-nose or heavy tweezer used to peel the unwanted cladding from the board. While peeling, use a soldering iron to soften the adhesive and heat the cladding to be removed.

PC board assembly

As I indicated earlier, the only component of this circuit designed to be surface mounted is the NE5205. The leads of the other components must therefore be bent to allow them to be mounted. **Figure 4** shows how to bend the leads of the various components so that they can be mounted. Try to follow this lead dress carefully since lead length is reasonably critical because of the high maximum frequencies of operation of this amplifier.

Although assembly of the PC board is straightforward, some care is necessary because of the tight packing of the components. You'll need a soldering iron with a 1/16th-inch tip to prevent solder from bridging between pads. Also, be very careful when mounting the five capacitors and the regulator. These are large components and their pads can be lifted very easily from the board when soldering. Both during and after mounting these components, take care not to bend them out of position; this would put considerable stress on the mounting pad and could cause the

Table 1. Component mounting sequence.

1.	NE5205	Pin 1 is at the Signetics "S".
2.	C3	Above the NE5205, solder ground plane.
3.	C4	Observe polarity, remember the bead, solder ground plane.
4.	R3	Solder ground plane.
5.	R4	Solder ground plane.
6.	CR1	Observe polarity, solder ground plane.
7.	C1	Observe polarity.
8.	C2	Observe polarity.
9.	R1	Solder ground plane.
10.	R6	Solder ground plane.
11.	R2	
12.	R5	
13.	R7	
14.	R8	
15.	R9	Solder ground plane.
16.	REG1	Observe mounting orientation.
17.	C5	Observe polarity, solder ground plane.
18.	C6	Observe polarity, solder ground plane.
19.	CR2	Observe polarity, leave anode lead full length.

pad to become separated, particularly after soldering. When soldering to the ground plane, keep the solder buildup to a minimum, since there's very little space under the board in the enclosure.

Figure 5 shows the assembly drawing of the PC board. Because it's probably the most difficult of the board components to mount — and because it will be partly covered by other components — mount the NE5205 first. The pin next to the "S" in Signetics is pin 1. The other components must be mounted in the order shown in **table 1** because of the tight packing. After mounting the amplifier, mount C3 over the top of the amplifier with its leads straddling the NE5205. Then mount the remaining components in the order shown in **table 1**, noting the associated comments.

When all the components have been mounted, bend four pieces of the resistor leads that were cut off during assembly into a "U" shape as shown in **fig. 4** and insert them from the ground plane side into the four pairs of holes at the corners of the board and solder them to the ground-plane side of the board. Make sure that the portion of each of these leads that's on the ground-plane side of the board is flush against the ground plane. They will be used to ground the board and hold it in the box. Finally, solder pieces of cut-off resistor leads to the input and output pads (again, bend as shown in **fig. 4**).

After the board is assembled, clean it in isopropyl alcohol; a 91 percent solution — probably available from your local drug store — works best. Don't use denatured ethanol because the denaturing agent isn't known. I prefer alcohol to trichloroethane for PC board cleaning because it presents no known health hazard. But it is *flammable*, so take appropriate precautions.

Figure 6 shows the assembled PC board before mounting in the enclosure.

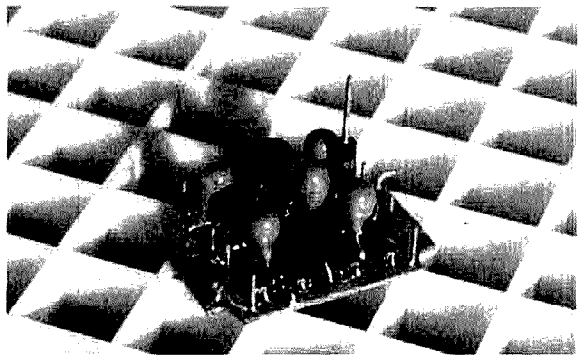


fig. 6. Assembled PC board.

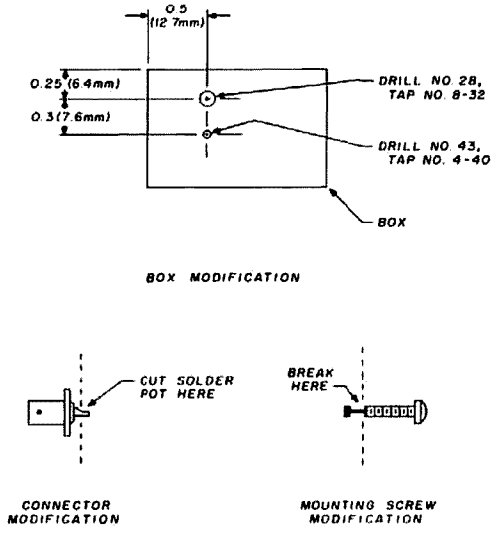


fig. 7. Box and connector modifications.

enclosure modifications

Enclosures for small RF projects are always a problem. The one I used for this amplifier was a small die-cast chassis box. This is a reasonably convenient enclosure; only two additional holes must be added for the power entry filter and the ground terminal. **Figure 7** shows the details of the modifications to the box, the two RF connectors, and the mounting screws. A word of caution here: don't try to cut the screws with a pair of diagonal cutters; the screws are very hard and trying to cut them will damage the cutters. The tips of the screws will break off easily enough if you use a pair of pliers to hold the tip while you bend the screw body with your fingers. Carefully lay out the holes using a precision scale and scribe. Center-punch the hole positions before drilling to prevent the drill from "walking." Tap the holes carefully; the tap is very brittle and will break with only slight side pressure.

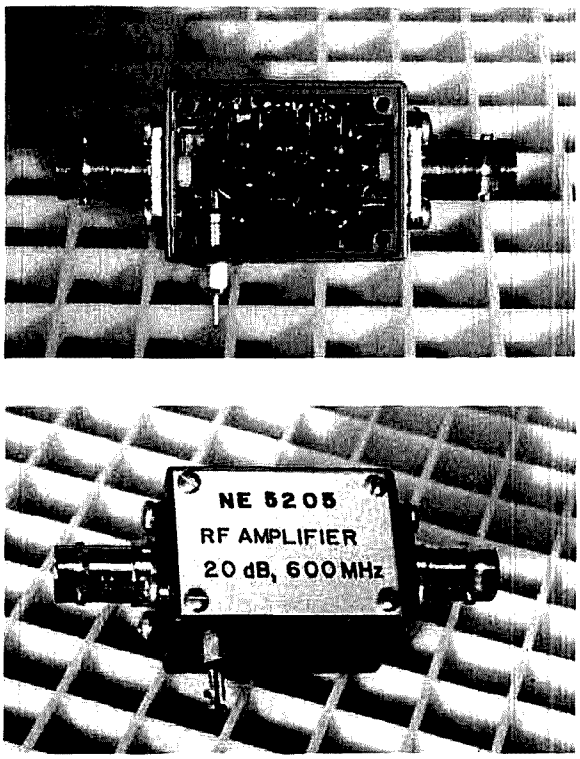


fig. 8. Completed amplifier assembly. A) Internal view. B) with cover on.

While tapping, use plenty of oil on the tap to prevent binding. Every turn or two, back the tap out to clear the chips. This will also help prevent binding.

final assembly

Now for the final task of putting it all together. With the connectors and power filter removed from the box, place the assembled PC board into the box, component side up and oriented so that the regulator is at the side of the box with the power entry filter. Loosely mount the input and output connectors, using only the upper two holes and the two short screws provided with the enclosure. As you place the connectors on the enclosure, be sure that the input and output connecting leads fit inside the connector holes. Now insert the longer screws into the bottom holes and make sure that they're on the component side of the PC board and between the pairs of leads coming up from the ground plane. Tighten all the screws. Now, bend the pairs of leads near each lower screw *over* the screw, laying them in the relief at the end of the screw. Solder all four screws to the wires and PC board. You'll need a larger soldering iron than you used for the board assembly to get enough heat for a good solder joint. Make sure that the tip of each screw has a good solder bead to the PC board; the PC board is grounded to the box only at these four places. Try to avoid getting solder in the screw threads so that the screws can be removed if you have to remove the board for repair later on. Solder the input and output leads to the connectors. Mount the ground terminal and power entry filter. If you cannot find a ground terminal, use a 4-40 brass screw about 1/2 inch long. Place a nut on the screw and screw it into the box until it's just penetrating the inside by a thread or two. Then tighten the nut down against the box to lock the screw in place. Place the bead, FB2, over the filter lead and attach the lead of CR2 to the power filter. Hold the body of CR2 with needle-nose pliers while bending its lead to prevent stressing its PC board mounting pad. Trim off the excess filter and diode lead.

Do a final cleaning by filling the assembled box with alcohol and letting it sit for a few minutes. Place your hand over the open top and shake the unit to stir the alcohol. Pour it out. Rinse with a little more alcohol and pour that out. Let the alcohol evaporate for a few minutes and then fasten the cover.

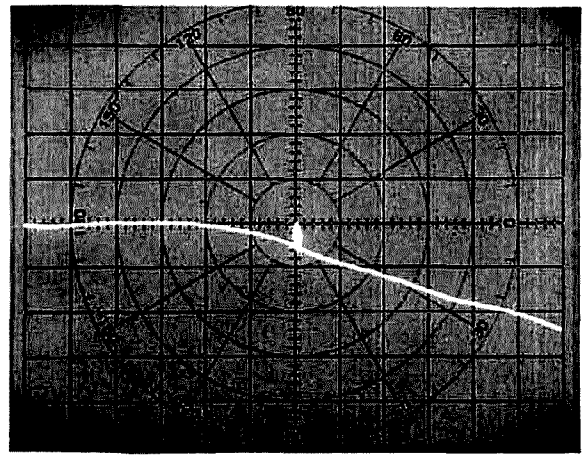


fig. 9. Amplifier frequency response.

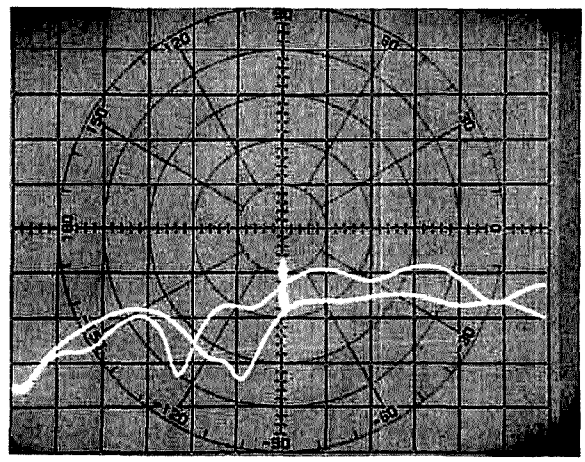


fig. 10. Input (S11) and output (S22) S-parameters.

Figure 8A shows the completed amplifier assembly with the cover removed; **fig. 8B** shows the completed assembly.

performance

Now for the proof test: how well does the amplifier work? If you followed the assembly instructions carefully, it should work rather well. **Figure 9** shows the frequency response of the unit that I built. The gain was 20 dB with a lower 3-dB point of 270 Hz and an upper of 608 MHz with usable gain to beyond 1.2 GHz. The bandpass flatness in the bandpass was better than ± 0.5 dB. The wideband noise figure was about 5 dB — not an LNA but certainly quite good. The output power at the 1-dB compression point at 100 MHz was +6.7 dBm, and the compression was reasonably constant with frequency up to about 600 MHz (the upper 3-dB cutoff frequency). The total supply current was about 35 mA and the minimum operating potential about 10 V. **Figure 10** shows the S-parameters S11 and S22 for the completed unit. These parameters are a ratio of the forward and reflected power at a given port. In **fig. 10** S11 is below about −20 dB up to about 500 MHz. The S11 parameter in this measurement was made at the input port. The −20 dB value shows that the power reflected back from the input is 20 dB below the power incident at the input port. That implies that the input impedance is reasonably close to the 50-ohm impedance of the test system; no power is reflected from a perfectly matched load. With a 20-dB return loss, the input impedance is within about 0.5 ohm of the characteristic 50-ohm impedance. S22 is the output port measurement. Up to about 500 MHz the reflected power from the output port is more than about 20 dB below the incident power at the output port. In this S22 measurement, power is actually applied to the output and the reflected power measured. So the output impedance is also reasonably close to 50 ohms (within about 0.5 ohm).

conclusion

This amplifier, if carefully constructed, provides very good performance for general RF experiments and RF projects where a relatively low cost, stable, low-noise, wide-bandwidth gain block is needed. The characteristics of the NE5205 make it a good choice for this application and the recommended circuit design assures both RF and DC stability as well as protection from typical experimental abuse. The recommended packaging provides good RF shielding and isolation, assuring stable, noise-free operation in most common RF environments. It is also relatively easy to construct. This amplifier should be a very useful addition to your collection of general RF amplfier designs.

references

1. Signetics NE5205 data sheet, January, 1985, Signetics Corporation, 811 East Arques Avenue, P.O. Box 3409, Sunnyvale, California 94088-3409.
2. Newark Electronics, 277 Fairfield Road, Fairfield, New Jersey 07006.
3. Amidon Associates, Incorporated, 12033 Otsego Street, North Hollywood, California 91607.

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ham radio TECHNIQUES

Bill Orr W6SAI

the grounded-grid amplifier

During the last decade a quiet revolution has taken place in Amateur Radio — the vacuum tube has virtually disappeared from the ham shack. I can't think of a new Amateur receiver, exciter, or transceiver using vacuum tubes that's sold in today's market.

But it's a different story in high-power, high-frequency amplifiers. In these, the vacuum tube remains supreme, in spite of several attempts to market a solid-state kilowatt amplifier. Such a device simply isn't cost-effective; I doubt that a practical 2-kW (so-called) solid-state linear amplifier will be available at a modest cost in the near future.

This leaves the power tube as the available high-power device. Today it's primarily a high-μ triode in a cathode-driven circuit (**fig. 1**) that's the popular choice for high-frequency SSB and CW service.

This circuit is ideal for Amateur service. It has good power gain and usually requires no neutralization in the HF region. Furthermore, it's hard to overdrive, and a certain portion of the drive power shows up as "free" power in the output circuit. The linearity of the grounded-grid amplifier is quite good and, all in all, it's a hard act to beat.

The circuit may be operated either as an amplifier or an oscillator depending on tuning. The tube grid is at (or near) RF ground potential and the driving signal is applied between cathode and ground. In amplifier service, when the cathode is driven positive by the exciting signal (with respect to the grid), the plate becomes more positive with respect to the cathode and also with respect to ground. In effect, the instantaneous plate voltage is developed in series and in phase with the exciting signal voltage. The driver and amplifier may then be considered as operating in series delivering power to the load (see **fig. 2**).

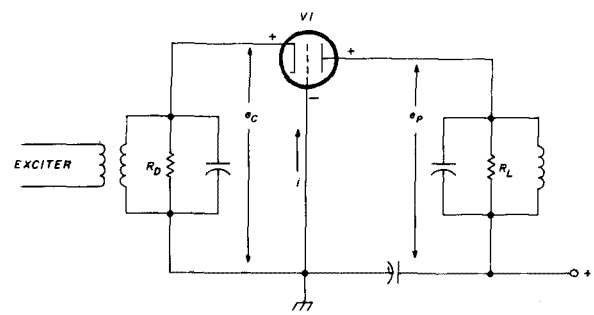


fig. 1. Simplified cathode driven circuit with R_D representing driver load and R_L amplifier load. Polarity of instantaneous RF voltages is indicated.

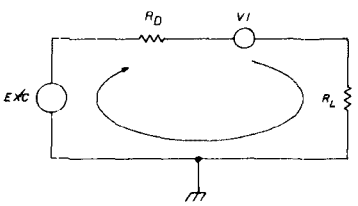


fig. 2. Exciter and amplifier are in series with respect to RF voltages. RF current flows through driver, as indicated by arrow.

In the better-designed cathode-driven amplifiers, a tuned circuit is used in the cathode to improve the regulation of the driver, to provide proper termination of the driver over the operating cycle, and to complete the plate circuit RF return path to the cathode of the amplifier. If the tuned cathode circuit is omitted, the various tasks fall upon the output circuit of the exciter. Many solid-state exciters cannot stand this set of operating conditions and may exhibit instability and undesired oscillation. The operator may jump to the conclusion that the amplifier is oscillating even though the problem is really in the exciter.

neutralize the grounded ground amplifier?

In the HF region, most grounded-grid amplifiers don't require neutralization because the feedback path from plate to cathode is small. However, the feedback path *does* exist and some Amateurs have discovered that the grounded-grid amplifier can become unstable and tricky to tune, especially on 10 meters and above.

It's easy to determine the degree of unwanted feedback in your amplifier. When fully loaded and tuned with carrier injection, maximum power output, minimum plate current, and maximum grid current should all coincide at one setting of the plate tuning capacitor. What? This doesn't happen in your amplifier? Maximum power output and

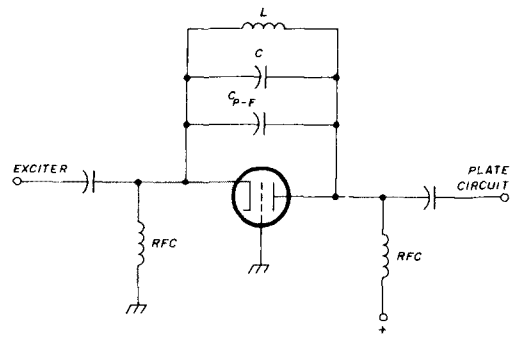


fig. 3. Inductive neutralization of cathode-plate feedback circuit. Resonant circuit is tuned to operating frequency.

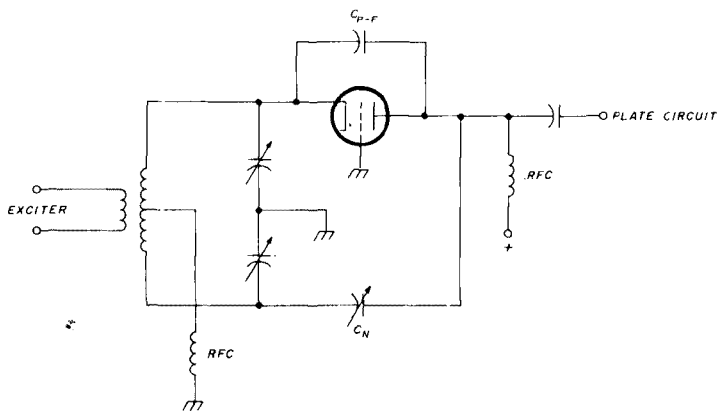


fig. 4. Bridge neutralization of cathode-plate feedback circuit.

minimum plate current don't coincide? You have some unwanted feedback in the amplifier. If the amplifier seems stable, don't worry about it.

the neutralizing circuit

Voltage feedback through the interelectrode capacitances of the tube from output to input can have a deleterious effect on amplifier performance under some circumstances. Control of this feedback is termed "neutralization." The purpose of neutralization is to make the input and output circuits of the amplifier relatively independent of each other during operation. It sounds simple, and it is. A balancing circuit (**fig. 3**), in which the capacitive feedback path is neutralized by making the capacitance part of a high-impedance, parallel-tuned circuit, is used. Another scheme takes a small portion of the RF plate voltage and feeds it back out-of-phase with the input voltage (**fig. 4**). When the value of the neutralizing capacitor C_N is approximately equal to the feedback capacitance of the tube, the circuit is balanced and will remain balanced over a considerable operating range of the amplifier.

VHF neutralization

In the VHF region (above 30 MHz) a second feedback path must be considered in amplifier operation. This path involves the grid-to-plate capacitance, the cathode-grid capacitance, and the grid lead inductance (**fig. 5**). Because the grid isn't truly at ground, because of the inherent grid lead inductance and other factors, a voltage may appear on the grid which can either increase or decrease the driving voltage. With sufficient unwanted grid voltage of the proper phase, the cathode-driven stage may oscillate, even though the tube has been neutralized.

There is, however, a certain frequency at which the two feedback paths tend to be self-cancelling. This is termed the "self-neutralizing frequency" of the tube and is usually found in the lower portion of the VHF band. This frequency is determined by physical tube size and the internal length of the conductors and elements within the tube.

Below the self-neutralizing frequency, the tube can be neutralized by the addition of a small inductance in the grid-to-ground path (**fig. 6**). Above this frequency, the tube can be neutralized by the addition of a series capacitance (**fig. 7**). Each of these neutralizing circuits is frequency-sensitive and the circuit must be adjusted if an appreciable change in operating frequency is made.

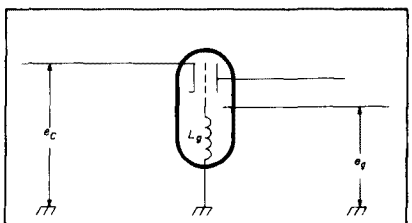


fig. 5. Voltage (e_g) develops across internal grid lead inductance which can increase or decrease driving voltage, e_g.

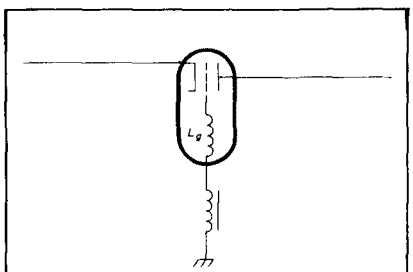


fig. 6. Below self-neutralizing frequency of tube, series inductor shifts point of self neutralization to operating frequency.

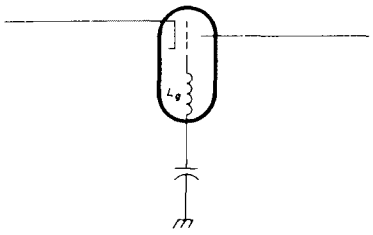


fig. 7. Above self-neutralizing frequency of tube, series capacitor shifts point of self neutralization to operating frequency.

a docile beast

In general, neutralization isn't required in Amateur amplifiers using cathode-driven circuitry below 30 MHz. The cathode-driven amplifier is a docile beast, with relatively low power gain when triode tubes are used and amplifier shielding is adequate. Amplifier instability at the operating frequency can often be cured by careful attention to feedback paths external to the amplifier (proper bypassing of primary power leads) and by ensuring that the exciter and amplifier are operating at the same ground potential. An extra-short, heavy ground strap between exciter and amplifier will often cure an unstable amplifier.

amplifier parasitics

Much has been written about amplifier parasitics. Some of it is true. As I said before, the cathode-driven amplifier is docile, and parasitics, when they occur, are usually mild (amplifier efficiency when oscillating in a parasitic mode is very low) and commonly above the self-neutralizing frequency of the tube.

A sure-fire cure for a parasitic is to load the circuit at the parasitic frequency until the amplifier refuses to oscillate. The tube lead common to all parasitic circuits is the plate; this is where parasitic suppression should take place (**fig. 8**). A simple resistor-inductor circuit will do the job. The inductor places the resistor across an appreciable portion of the plate lead at the parasitic frequency and thus loads the circuit. At the operating frequency, the resistor is across only a small portion of the electrical length of the plate lead and is almost "invisible." Too many turns in the inductor will couple the resistor too tightly at the operating frequency and the resistor will dissipate a portion of the amplifier's fundamental power and will probably overheat. If the parasitic is truly suppressed, then there will be no parasitic power. Too few turns in the inductor and the suppressor won't do its job. Cut-and-try is the keynote to success in this operation.

the "Rocky Point" effect

The vacuum in a modern power tube is on the order of 10 Torr (millimeters of mercury) in order to maintain proper cathode (filament) emission and to provide adequate insulation between the electrodes. In spite of the high vacuum employed, it sometimes happens that the insulation between the anode and other electrodes suddenly breaks down, with flash-over occurring inside the tube.

This phenomenon has been referred to as the "Rocky Point" effect, after the RCA transmitting site where it was first observed in the 1930s. The effect can occur at low voltages and in relatively small tubes. Though it's not attributable to a gradual deterioration of the vacuum, it can be brought about by a small quantity of ions liberated from an electrode. Ionization causes an electrical discharge, whereupon the ions disappear, either because of absorption or action of the getter within the tube. If the equipment is power-supply limited, the discharge usually causes no damage. However, most Amateur equipment makes use of a high-capacitance filter in the power supply, and the energy of this capacitor is "dumped" into the flash-over. Since the tube, at that instant, forms a virtual short circuit, the discharge current can run very high, damaging the tube electrodes, metering circuits, and associated components.

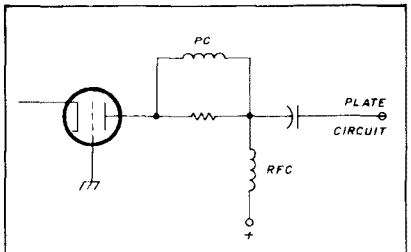


fig. 8. Proper position of parasitic choke is in the plate lead of cathode driven amplifier.

The flashover is very sensitive to a drop in plate potential. The easiest way to control it is to insert a small series resistor in the plate supply (**fig. 9**). The voltage drop across the resistor during a flashover will lower the plate voltage and extinguish the vacuum arc.

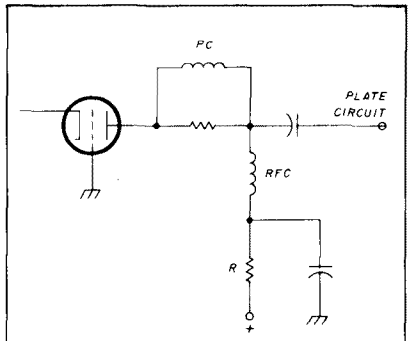


fig. 9. Resistor in series with B-plus protects tube and circuitry in case of flash-over.

Without the resistor, substantial damage may occur to tube and amplifier components. With the resistor in place, in the case of a rare flash-over, the operator will be aware only of a soft "pop" or "snap" and amplifier operation will continue as before. The resistor limits the current while the energy in the filler capacitor is being dissipated.

In most cases a 50-ohm, 20-watt resistor incorporated in the B-plus lead after the filter capacitor (either in the amplifier or in the power supply) should provide adequate protection. Inexpensive and easy to install, the resistor can protect a power tube worth many hundreds of dollars.

The resistor is in a high-voltage circuit and should be adequately insulated from ground. Mounting it on ceramic insulating pillars is a good idea.

filament voltage

It's a good idea to check your filament

voltage at the socket in a cathode-driven amplifier. A voltage drop may exist across the filament choke and thus reduce available filament voltage at the tube. The voltage should be checked with an RMS-responding meter (iron-vane type, for example) and not with the garden-variety volt-ohmmeter which, more often than not, employs a DC meter and a series-connected diode rectifier to measure AC voltages. This combination is often inaccurate as the diode ages and its response to the rough waveform of the common AC primary line is questionable.

Filament voltage should be held to the tube manufacturer's specification limits, generally ±5 percent of the designated voltage. For Amateur service, it is generally prudent to remain on the low side of the voltage limit, rather than on the high side. (I generally run my tube filaments about 2 percent below the suggested operating voltage.)

RF feedback

In some cathode-driven amplifiers, a small degree of negative RF feedback is incorporated. This absorbs some excess drive power, tends to make the amplifier more stable, and improves the intermodulation distortion figure slightly. The feedback circuit is made up of the plate-to-grid capacitance, which is set by the equipment manufacturer. A representative circuit is shown in **fig. 10**. Feedback is about 2 to 3 dB in the case illustrated. Decreasing the grid-to-ground capacitance raises the feedback level, but also tends to degrade the grid-filament isolation at the operating frequency. A happy compromise must be found for the circuit to do its job. In the case of an amplifier using two 3-500Z high-μ power triodes, 600 pF seems to be satisfactory. This is accomplished by placing a 200-pF capacitor at each grid pin to ground. The capacitors are shunted by an RF choke to complete the DC ground return for grid current. The choke has nothing to do with the operation of the RF circuit.

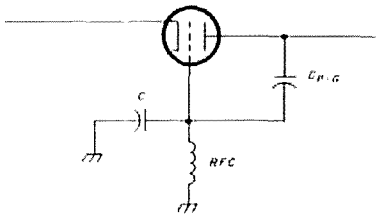


fig. 10. Plate-to-grid capacitance plus capacitor C forms voltage divider to impress negative feedback voltage with respect to cathode. RF choke completes grid DC return.

summary

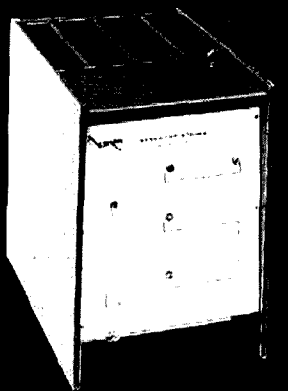
The grounded-grid (cathode driven) configuration is admirably suited to Amateur service in the HF and VHF regions and the circuit performs well in a properly designed and operated amplifier.

the EME directory

A second printing of the *EME 144 MHz Directory* lists active moonbounce stations, their addresses and locations, equipment used, and other pertinent information. For a copy, send five first-class stamps (or IRCs) to me at: Varian EIMAC, 301 Industrial Way, San Carlos, California 94070.

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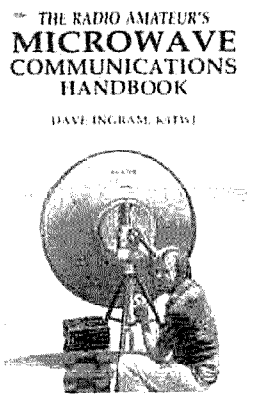
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remotely controlled stations:
a look at a successful remote base

Stuck in an RF gulch? Try another site

Imagine having a "big gun" 160-meter station — but with no antenna consuming your back yard. Imagine running 1 kW on 6 meters — with no fear of TVI. Imagine DXing into the next state on 1296 MHz — but the only rig at your QTH is a 450-MHz handheld! All this is possible using a remotely controlled station, commonly known as a "Remote Base" or simply a "Remote."

A Remote may consist of as little as a simple wireline control link operating a single-channel 2-meter rig, or it may be a complex array of computer-controlled gear, covering HF through microwave and controlled via 450 MHz-FM. Ideally, it's positioned on top of a mountain or tall building, like a repeater, but it could be located anywhere — even in the trunk of a car!

Possibly the biggest advantage of a Remote is that it allows antennas to be located at a better site than may be available at home. Many housing developments, for example, don't allow outside antennas. And even where they're allowed, TVI may prohibit HF or 6-meter operation at appropriate transmitter power levels. Or perhaps you just need to get your antennas out of the "RF gulch" you're trapped in.

But a better antenna location isn't the only reason to go Remote. A Remote allows several hams to consolidate their resources into one superior station, accessible to all. A group of operators then needs to purchase only one HF rig, one 1296-MHz station, and a single antenna farm instead of duplicating their efforts individually at great expense.

The sharing of talents is valuable, too. In any group of hams there are likely to be those with specialized knowledge of RF, digital communications, and antennas, for example. Working together and pooling resources, each can contribute his or her skills toward building a station far superior to what each might accomplish separately.

Located on a 2000-foot mountain near Ventura, W6ORE Remote (**fig. 1**), the station described in this article, has been operating successfully for ten years. Using a 450-MHz FM control channel, this Remote allows the users to operate an HF rig covering 160 to 10 meters and an FM VHF/UHF station covering 144 to 148 MHz, 220 to 225 MHz, and 440 to 450 MHz. Also controlled are an X-band beacon and a programmable speech synthesizer, along with other miscellaneous functions.

Every one of these items can be controlled from the user's home QTH or automobile (or boat, in one case), with a rig as simple as a handheld equipped with a rubber duck antenna and a touchtone pad.

The HF rig can be set directly to any frequency, or scanned or stepped up or down the bands in 100-Hz steps. A speech synthesizer reads out the frequency and reports if the transmitter is becoming overheated, or if a band edge has been reached while scanning, and to what portion of the band (Novice, General, Advanced, or Extra) the operator is tuned.

The VHF/UHF station, a Drake UV-3, can also be directly set to any frequency, scanned, or stepped up or down. Repeater splits are set automatically, but any odd split, even cross-band, is possible. All of this is controlled, "on-the-fly" with just a few keystrokes on a touchtone pad!

the computer

As you may have guessed, a computer (**fig. 2**) controls the system. Indeed, the heart of this Remote is a custom designed Z80™ computer built and programmed by Bob Schellhorn, W6ORE. The Z80 CPU was chosen for its rich instruction set and extensive I/O (Input/Output) capacity.

By Steve J. Noll, WA6EJO, 1288 Winford Avenue, Ventura, California 93004-2504

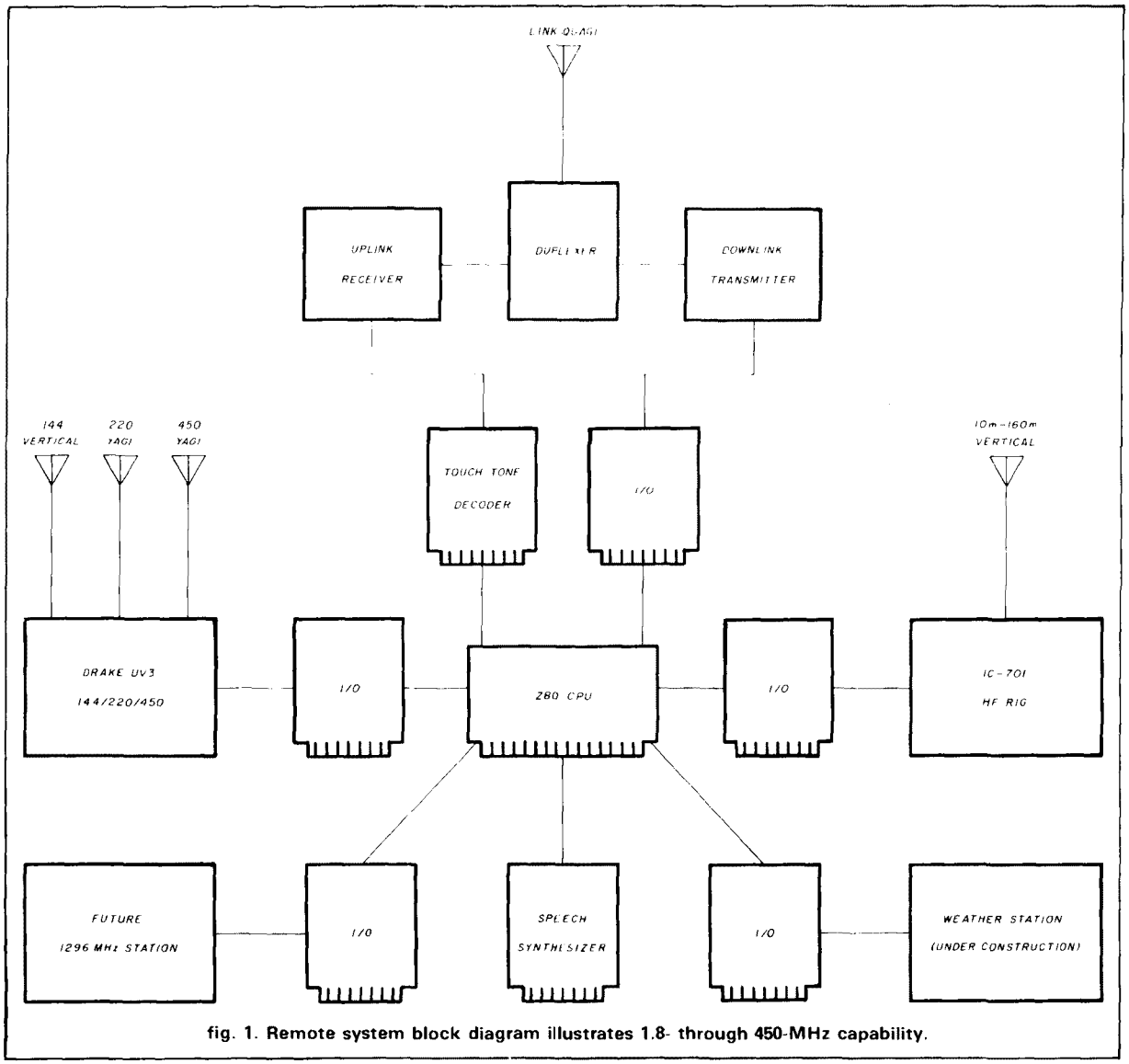


fig. 1. Remote system block diagram illustrates 1.8- through 450-MHz capability.

The software was written entirely in assembly language for speed and efficient use of memory. The software, known as "firmware" at this point, is contained in a 27128 UV EPROM (ultra violet light erasable, programmable, read-only memory). Transient and changeable information is stored in a 6264 CMOS RAM (complementary metal oxide semiconductor random access memory).

The Z80 and memory are built on an S-100 breadboard, although the S-100 bus protocol isn't used. S-100 cards afford generous wiring space and a 100-pin edge connector provides for plenty of I/O and allows the card to be easily removed for modifications (see **fig. 3**).

The CPU card is enclosed in an RF-tight box fashioned from un-etched copper-clad printed circuit board. Custom RF-tight enclosures constructed with surplus printed circuit board are often cheaper than store-bought aluminum chassis, which also never seem to be available in just the right size. An RF-tight CPU box is a must, as computers are notorious for generating excessive radio "trash." And in this case, the computer is mounted within inches of several radio receivers.

The Z80 runs at a fairly low speed to minimize RFI. The I/O is further slowed by a combination of special software and hardware techniques and brought out of the CPU box using feedthrough capacitors. A card cage receives the CPU I/O lines and distributes them to various 5 x 7-inch I/O cards. The card cage isn't shielded, but because the I/O is slowed at this point, no noticeable RFI is generated.

The I/O cards provide the digital and audio interfaces to the HF rig, the VHF/UHF rig, the 450-MHz control up-



fig. 2. A large double cabinet provides room for future expansion.

link receiver and downlink transmitter. One card contains the speech synthesizer, which is so necessary to the operation of such a complex system.

a speech synthesizer is a must

The speech synthesizer is built around the Votrax SC--01 speech chip. This is a phoneme-based chip, which makes it completely user-programmable. It doesn't provide the latest, state-of-the-art high-quality speech; in fact, it sounds rather like a robot. However, it's far superior to many other speech chips in that it can say absolutely *anything*. No compromises (such as spelling out words not in the limited vocabulary common to other synthesizers) have to be made. It's also very memory-efficient. Only about five phonemes, or five bytes of memory, are required to store an average word.

the HF station

The HF rig (**figs. 4** and **5**) is an ICOM IC-701, a model no longer in production but a natural for computer control because of its 24-pin connector for plugging in a remote control unit. Through this connector the computer can set frequencies, scan or step in 100-Hz increments, and transmit. The computer keeps track of the current frequency, the band edges and the license class boundaries. Of course, frequency can be read at will (a single touchtone on the uplink) and is announced politely on the downlink – i.e., "Fourteen-point-three-one-four."

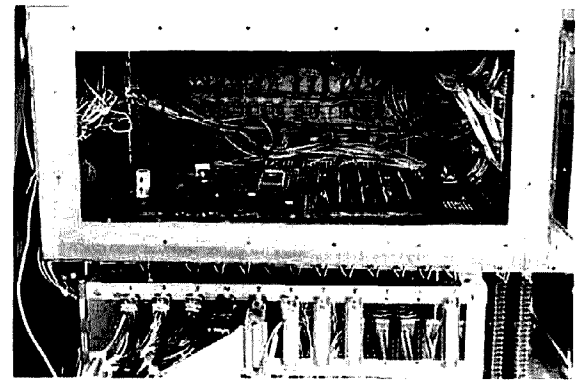


fig. 3. Inside of computer controller. Much of the visible wiring is for the diagnostic front panel.

All is not perfect, though. Outputs weren't provided for mode switching (USB/LSB) or for over-temperature warning. Minor surgery provided these needed outputs – two, in fact, for the over-temperature alert. When the rig's fan comes on, the speech synthesizer announces "Overheat warning" on the 450-MHz downlink. If the PA temperature continues to rise, the transmitter shuts down and the synthesizer announces, "Overheat shutdown."

A ground-mounted Butternut vertical allows 160- to 10-meter coverage without an antenna tuner. Because the Remote is located on cattle grazing land, a small corral was built around the vertical to ward off any itching Elsies. A 160-meter wire antenna is planned for the future; there will be no problem in having the computer switch over automatically when a 160-meter frequency is selected.

the VHF/UHF station

Another out-of-production rig, a Drake UV3, provides the 144/220/450-MHz coverage. This rig was designed with a removable control head to allow the RF section to be located in the trunk of a car, while the head is mounted under the dash – again, a wonderful opportunity for easy computer control. The removable head is replaced by a homebrew interface box allowing the computer full access to frequency setting and band and power level changing.

The VHF/UHF antennas are a simple ground plane on 2 meters and short KLM log-feed Yagis for 220 and 450 MHz. Luckily, most of the action is in one direction from the Remote site so that directional (gain) antennas can be used. The short log-feed Yagis afford wide bandwidth and a not-too-narrow beam shape (see **fig. 7**).

the control link

Primary to the operation of the Remote is the 450-MHz uplink, which allows controlling and talking through the

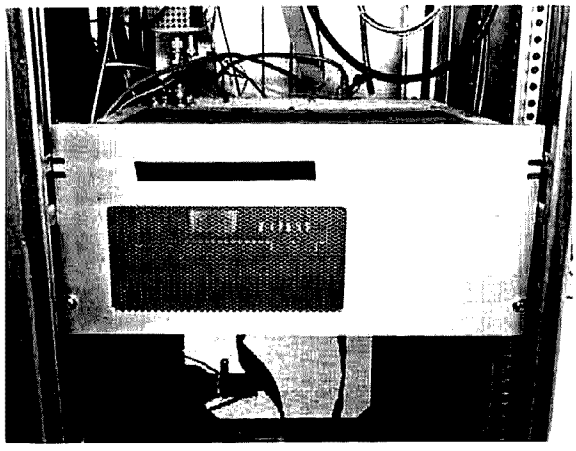


fig. 4. HF rig is in a shielded rack-mount box.

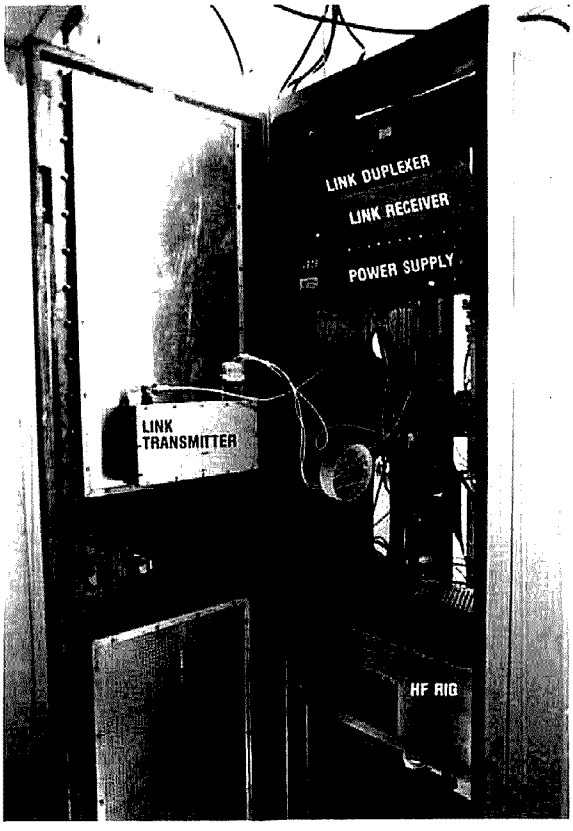


fig. 5. Rear view of cabinet illustrates placement of link components.

HF and VHF/UHF stations. The 450-MHz downlink pipes the HF/VHF/UHF receive audio back to the user.

The uplink receiver is a Yaesu 708R, somewhat modified to accommodate a Motorola squelch chip. The downlink transmitter is a cannibalized ICOM IC-30. Each connect through a four-cavity Phelps Dodge duplexer to a homebrew Quagi. All of the users live and work in the same general direction from the Remote site, which permits the luxury of using a high-gain directional link antenna.

some legal points

The Remote control requirements of Part 97 are often violated or ignored by remotely controlled station users. Because the regulations, which are quite complex and often confusing, require careful study for compliance, some definitions are in order:

- **Auxiliary operation:** *radio communication for remotely controlling other amateur radio stations* (97.31). *All amateur frequency bands above 220.5 MHz, except 431-433 MHz and 435-438 MHz, are available for auxiliary operation* (97.86d).
- **Remote control:** *manual control, with the control operator monitoring the operation on duty at a control point located elsewhere than at the station transmitter, such that the associated operating adjustments are accessible through a control link* (97.3m2).
- **Control link:** *apparatus for effecting remote control*

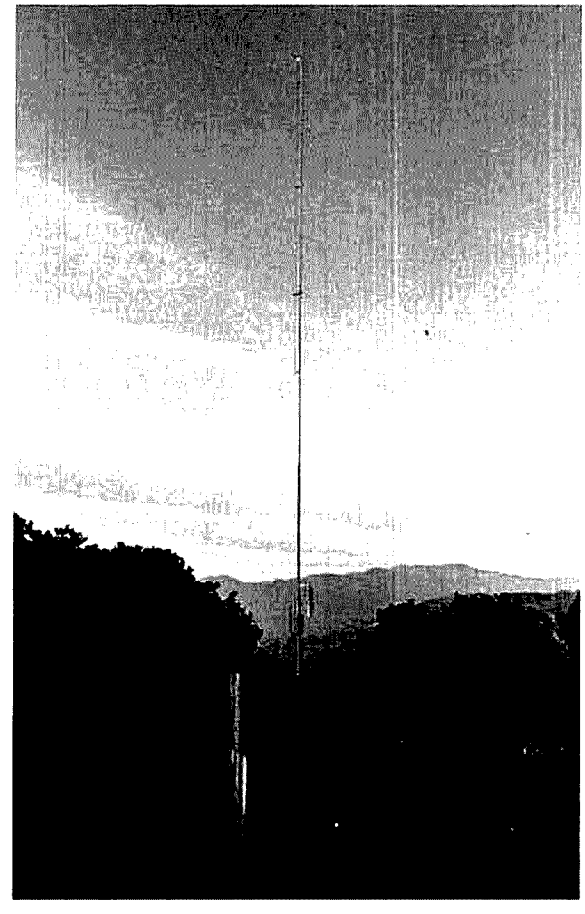


fig. 6. The HF vertical is protected from cattle and deer by a small corral.

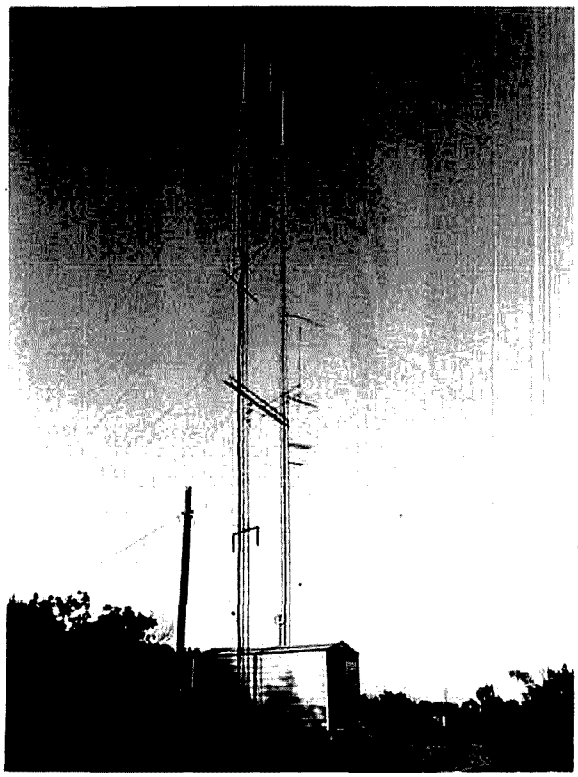


fig. 7. The remote site shares towers and building with commercial services. HF vertical is visible at bottom right.

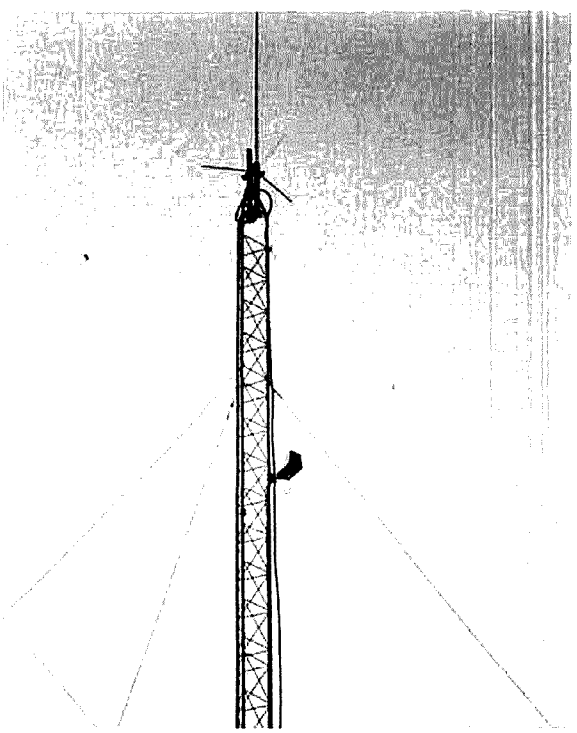


fig. 8. X-band beacon is perched near top of tower.

between a control point and remotely controlled station (97.3n).

Thus, a station can be remotely controlled through a control link; this is considered auxiliary operation. Note that Part 97 never uses the popular terms "Remote" or "Remote Base."

Of course, *"The frequencies available for use by a control operator of an amateur station are dependent on the operator license classification of the control operator. . ."* (97.63c). Even if a Remote is licensed to an Extra, a General class control operator is still restricted to the General bands.

On identification: . . . *a station in auxiliary operation shall transmit the word "auxiliary" at the end of the station call sign* (97.84d2). So on the control uplink, I identify as "WA6EJO AUXILIARY." The control downlink identifies itself as "W6ORE AUXILIARY." When transmitting through the Remote's HF station, however, the only required ID is that of the remotely controlled station licensee.

On control link security: *A station in auxiliary operation shall be used only to communicate with stations shown in the system network diagram* (97.86d). To comply with this regulation, a non-member of a Remote breaking into the link should be informed that it is a control link and politely asked to leave.

On remotely controlled station security: *Each remotely controlled station shall be protected against unauthorized station operation, whether caused by activation of the control link, or otherwise* (97.88g). "Automatic control" is not advisable.

can a remote be used for gaining contest points?

Use of repeaters is usually prohibited for contesting, but a Remote is merely a station operated by remote control, not a repeater. The ARRL has affirmed this and does not differentiate between Remote and "regular" station contacts. Note that users of a given Remote can't make contest points under their own calls. The transmitter(s) of a Remote are licensed to the licensee of the Remote and therefore bear only his or her call sign, not each individual user's.

operating particulars

Nowadays, most 450-MHz rigs include built-in subaudible tones (PL). The ICOM IC-04AT is one example. All PL frequencies are built in and merely selected from the keyboard. The result is increased convenience but at the cost of reduced security. A remotely controlled station isn't very secure if left unattended with only PL protection. And Part 97 *requires* that Remotes be well protected.

For that reason the W6ORE Remote is left off when not in actual use. A user turns the Remote on by sending a touchtone code on the uplink and is responsible for

operation of the station until turned off, or until control is passed on to another user. On turn-on, and every 10 minutes thereafter, the synthesized voice IDs the downlink, "W6ORE Auxiliary." At this point, the user can enter modes for controlling the VHF/UHF station, the HF station, or miscellaneous functions.

If VHF/UHF station control is selected, any frequency that the UV3 can cover can be entered via a touchtone pad. Any split, normal, reverse, or odd can be set. Of course, all frequencies and split settings are announced on the downlink by the speech synthesizer. If the UV3 transmitter is enabled, a beep is sent on the downlink every time the user drops his or her uplink carrier to serve as a reminder that the transmission is occurring on another frequency.

Now the computer really proves its worth. Each user is assigned a "scan memory." Each scan memory holds 32 "channels," just like a regular scanner. A user can place any mixture of 2-meter, 220-MHz, and 450-MHz frequencies in his or her scan memory. After turning the Remote on, the user needs to send only a single touchtone on the uplink to activate the scan memory and the UV3 to begin scanning! Scanning, of course, stops on a "talking" channel, but channels can also be "locked out" temporarily or permanently with a keystroke.

Operation of the HF rig is similar. A frequency can be entered directly with touchtones. Then a touchtone code starts the IC-701 tuning up, or down, in 100-Hz steps.

Keying a mike on the uplink briefly will reverse the scan direction and slow it down. Keying the mike a little longer will stop the IC-701 from tuning. At any time a certain touchtone will cause the speech synthesizer to announce the current frequency being monitored. Another touchtone will single-step the rig in 100-Hz steps; yet another will toggle the rig between WWV and the last frequency.

Of course, none of these features are built into the HF rig. The computer controlling the Remote has added them. In addition to the user scan memories, the computer provides 90 single-channel "slots." Each slot can hold any VHF, UHF, or HF frequency. Simply entering a slot number turns on the appropriate rig and sets it to that frequency. For example, one slot holds the 20-meter Maritime Mobile net frequency, a few slots are assigned to local 2-meter repeaters, and one series of slots contains all of the HF W1AW voice bulletin frequencies.

HF apprehensions

There certainly were apprehensions about Remote operation of the HF rig before it was installed. What would tuning a rig several miles away be like without having a frequency display to watch or even a knob to grasp? Would we have to operate duplex, tuning on the uplink while listening on the downlink? The concept of blind operation and computer control was first tested by connecting the HF rig to the author's NorthStar Horizon S-100 computer. A program in BASIC took commands from the computer keyboard and sent them to the rig through parallel ports. It worked great. There were no problems in operating the rig without seeing its front panel. With a little practice a sideband signal could be scanned up to and stopped directly on.

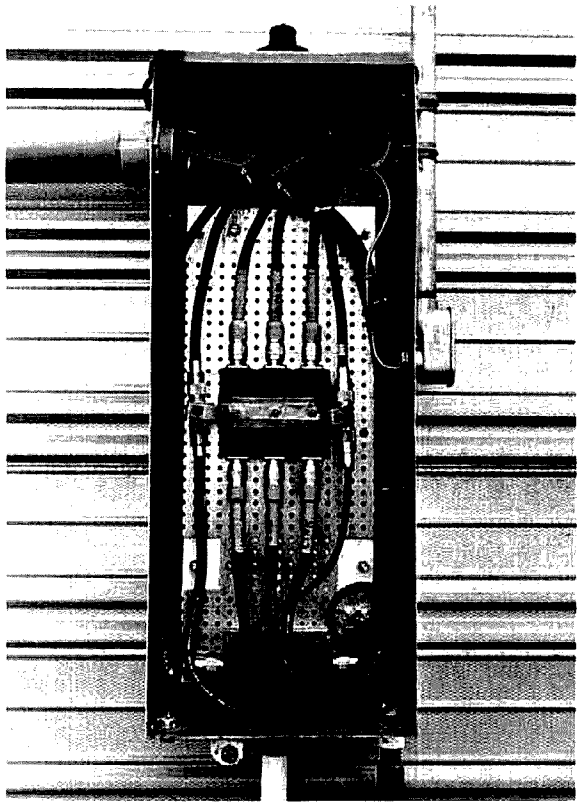


fig. 9. Coax passes through lightning arrestors before entering building.

There is a utility mode that allows users to control a multitude of items. Uplink squelch characteristics can be adjusted. The Remote's battery charger and a 140-mW X-band beacon can be controlled. (See **fig. 8.**) The speech synthesizer can be programmed on the phoneme level to say any word or message. Any portion of the EPROM memory can be read (the speech synthesizer reads out the locations in hex). The static RAM can be read and also written to. The speed that the HF rig tunes can be adjusted. These are just a few of the things that can be controlled.

keeping the commands simple

Even though there are several dozen different touchtone commands, the majority of the most used are only single digit. This makes life much easier. There are no vast tables of long commands to be memorized. This is achieved mainly by dividing major operating areas,

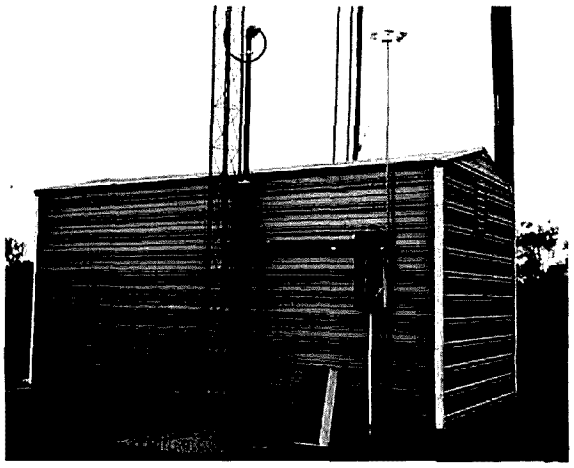


fig. 10. Coax enters building through a weather-proof box. The anemometer is made from a DC motor and recycled air freshener containers.

i.e., HF, VHF/UHF and utility, into separate "modes." The touchtone decoder has to be fairly good to keep uplink voice falsing down when single-digit codes are used. The Mitel MT8860/8865 chip set is used.

bells and whistles (figs. 9 and 10)

A weather station addition is now nearing completion. It is built around a National ADC 0817 16-channel analog to digital (A/D) conversion chip. The chip has 16 analog inputs and an 8-bit digital output. It is quite easy to connect to a computer. The computer needs to send the chip a 4-bit address to set the chip to read the desired channel. The chip then measures the voltage on that channel and presents the data in an 8-bit, 0-255 format. It is then up to the computer to multiply or divide this number by a constant to come up with a meaningful measurement. For example, the home-made anemometer generates 0.027 volt/mile/hour. An op amp amplifies this signal by 2.47. The A/D converter sensitivity is 0.02 volt/bit. Thus, if a 10 mph wind is blowing, the A/D will put out (10 x 0.027 x 2.47)/0.02 or 33 rounded. The computer then multiplies the A/D output by 0.303 to get 10 mph, which is announced by the speech synthesizer on command. Also measured are inside and outside temperature, humidity, AC line voltage, battery voltage, charger current, and HF antenna VSWR.

summary

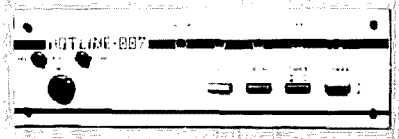
Building a remotely controlled station is unquestionably a major project, but the rewards are well worth the effort. Computer control and synthesized speech feedback are a must for all but the simplest systems. There is commercial hardware available for remote control, but you might want to consider "rolling your own" to gain the ultimate in versatility.

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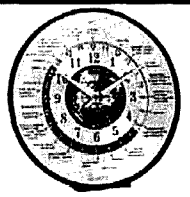
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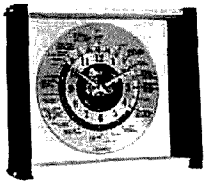


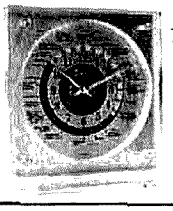


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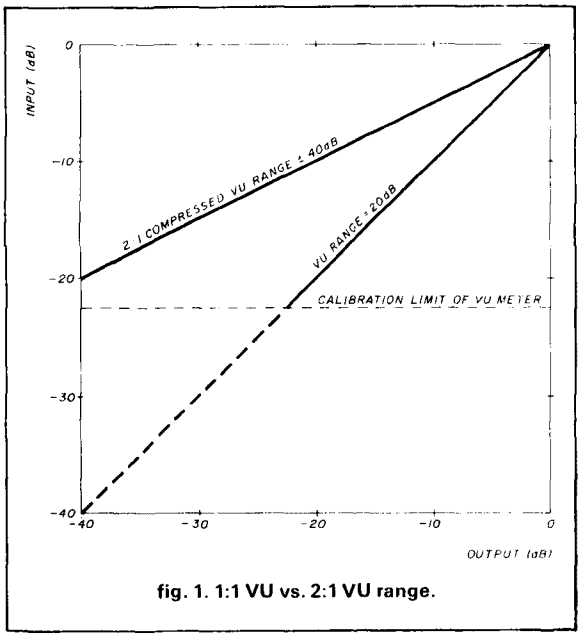


fig. 1. 1:1 VU vs. 2:1 VU range.

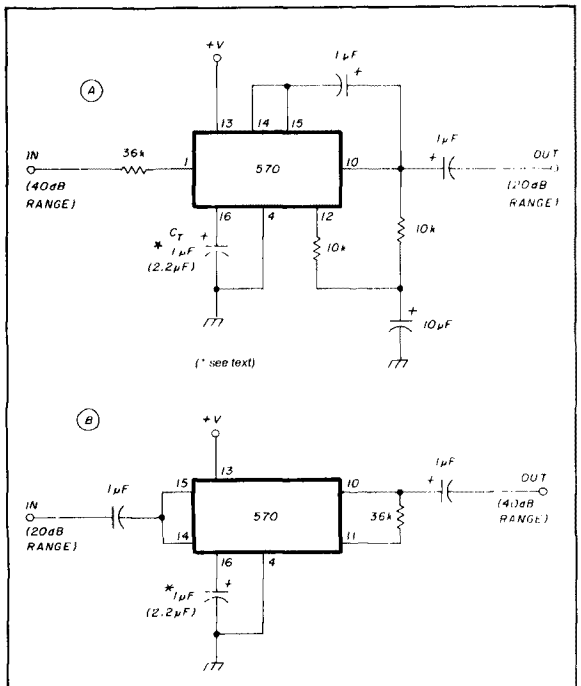


fig. 2. The NE570 used as (A) compressor and (B) 1:2 expander.

In Amateur Radio we're accustomed to signal-to-noise ratios of less than 15-20 dB. On UHF, however, there's sometimes a need to measure SNR higher than 20 dB. An example of this would be in setting up VHF or UHF networks or linking systems to minimize hum, noise, or PL leakage.

One method of extending the range of a standard VU meter from its usual 20-22 dB to 40-45 dB is to use an NE570 or NE571 2:1 compression amplifier between the source signal and the VU meter. This device provides a very accurate 2:1 curve so that accurate SNR readings can be achieved (see **fig. 1**).

circuit description

The NE570 consists of three basic blocks: a 741-like op amp, a gain control amplifier, and a precision detector. It can be set up to compress or expand at a 2:1 ratio, depending on how these elements are connected (see **fig. 2**).

The output of the NE570 is buffered by an MC3403 op amp to provide gain in the unit and better drive to the VU meter (the NE570 is limited in this regard). I chose the MC3403 because it has low crossover distortion and requires only a single voltage supply.

(The circuit board artwork from the Project OSCAR ACSSB Level 1 TX Adapter is shown in **fig. 3**). A layout for the VU Extender and its schematic are also included (**fig. 4**). You may wish to use the extra op amp sections to provide a standard audio weighting curve for some kinds of measurements. There are three op amps still available in the MC3403 to provide this function.

Calibration is accomplished by setting the input attenuator pot (R_A) in **fig. 4** to mid-scale, then adjusting R_B with the selector switch (S1) in the "OUT" position until you read "0" VU with 0.773 volts AC

By James Eagleson, WB6JNN, 15 Valdez Lane, Watsonville, California 95076

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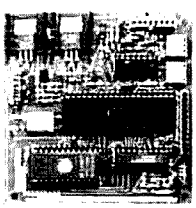
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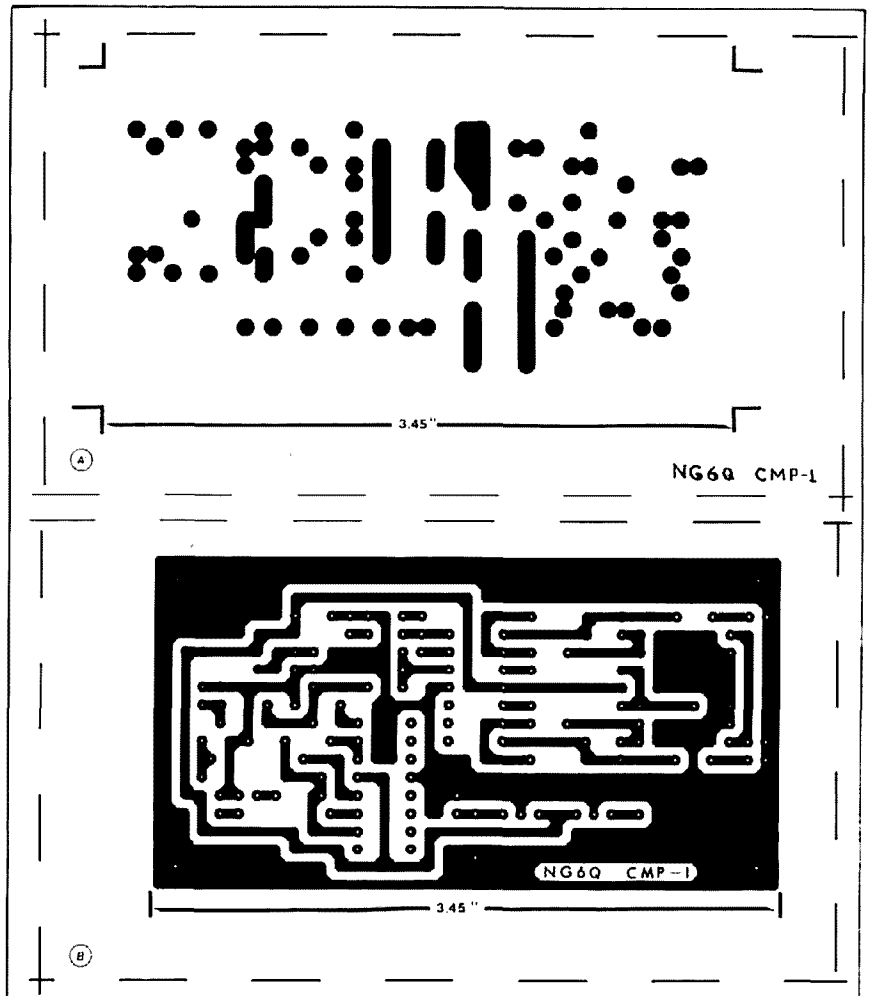
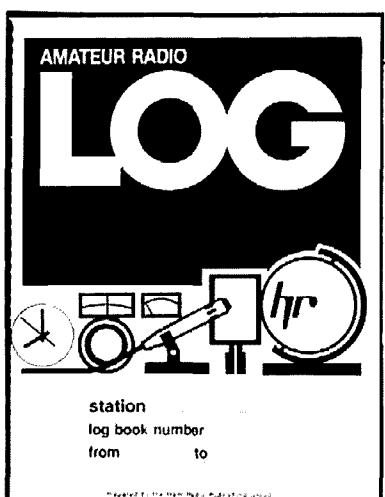


fig. 3. VU extension circuit printed circuit board: (A) component side; (B) solder or ground side. Note: if reader interest warrants, Project OSCAR will prepare printed circuit boards. Address inquiries to author (enclose SASE).



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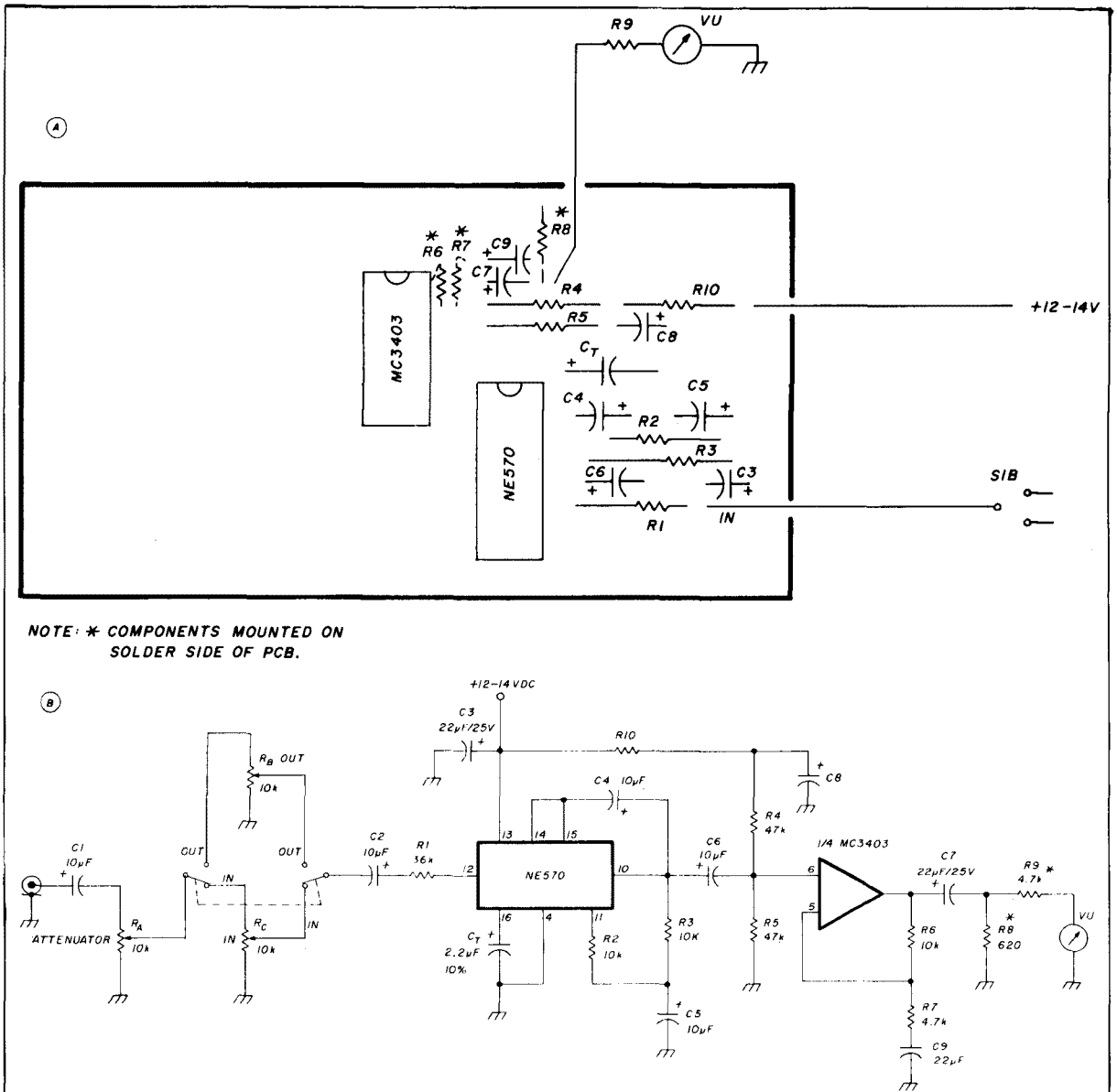


fig. 4. VU extension circuit printed circuit board: (A) component layout; (B) extension circuit.

at the input. This can be checked with a VOM or digital voltmeter. After switching S1 to "IN," adjust R_C for the same reading. You'll now notice that by dropping the input level 10 dB in the "OUT" position, you'll read −5 dB when you switch to the "IN" position.

All readings taken with the adapter "IN" should be multiplied by 2 (i.e., −5 dB = − 10 dB) to obtain actual SNR.

performance

The unit works well for normal speech and CW because the time constants have been chosen to complement the usual syllabic rate of voice and CW, which is about the same.

The 2.2 µF capacitor specified for (C_T) gives good performance down to about 100 Hz.

There will be some overshoot and undershoot on the system (no AGC amplifier is perfect), but it will certainly give results within a few dB of correct levels and costs far less than equivalent commercial units capable of measuring the same information. I've found it a useful addition to my shack.

ham radio

a very sensitive
LF or HF field-strength meter

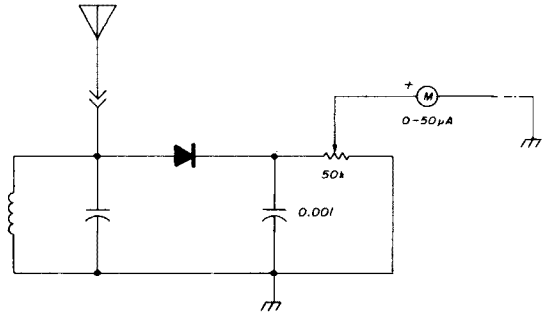


fig. 1. The typical field-strength meter circuit will work for HF medium power applications, but is useless for 1750-meter QRP testing.

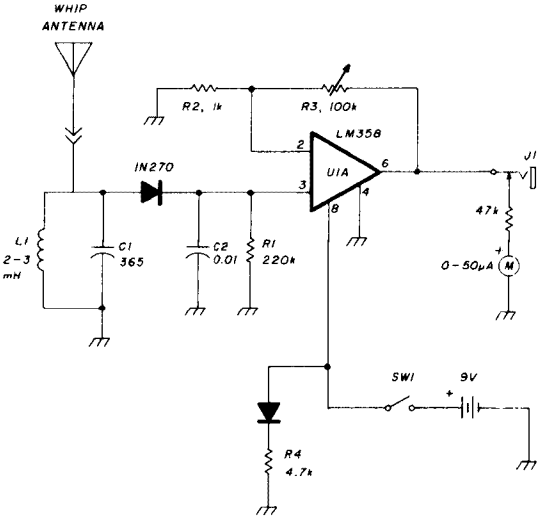


fig. 2. A single 9-volt battery will power the field strength meter, with a total current drain of 3 mA.

Detect small changes with this handy circuit

While the simple resonant tank, diode detector, and microammeter-type field strength meter (**fig. 1**) may be usable for HF signal evaluation, it's almost useless for any reasonable measurements on 1750 meters, where a 1-watt input restriction applies and very low ERP is the rule.

Using the standard diode detector scheme, I found that the usable scale readings for 1750-meter tests were limited to a maximum of 30 feet away from the antenna to be evaluated. To compensate for this insensitivity, a larger antenna connected to a field strength meter would be required. But this would make field-level measurements cumbersome. One solution to this problem would be a DC amplifier at the output of the detector to provide the gain required for driving the meter indicator.

circuit description

In **fig. 2** the complete LF field strength meter circuit is shown. C1 and L1 are made to resonate on the 1750-meter band, with the total coverage being from 150 kHz to 500 kHz. L1 can be slug-tuned for 160-to-190 kHz coverage alone or a 2.5-mH choke can be used for L1, if desired, using C1 for tuning. A 1N270 germanium diode rectifies the RF signal and C2 is charged at the peak RF level. This DC level is amplified by U1, an LM358 operational amplifier requiring only a single 9-volt supply for operation. The gain is determined by R2 and R3. R3 is a 100-kilohm linear potentiometer that varies the DC gain from 1 to 100, driving the 50-microampere meter, which acts as a voltmeter in conjunction with R5. A normally closed 3.5-mm jack is connected in series with the analog 50-microampere meter for remote meter readings and/or a DC level which can power an audio

By S.J. DeFrancesco, K1RGO, 17 Jeffry Road, East Haven, Connecticut 06512

Table 1.

L1	C1 (variable)	Frequency Range	Ham Band
50 μH	30-365 pF	1- 4 MHz	160, 80 meters
3 μH	30-365 pF	5-16 MHz	40, 30, 20 meters
0.9 μH	30-365 pF	9-30 MHz	30, 20, 15, 12, 10 meters
2.5 mH	—	Broadband at reduced gain	

oscillator for CW sidetone operation. An LED was added to indicate the "on" status.

You can expect long battery life because the amplifier will continue to operate even when the battery voltage drops to 4 volts. With 9 volts applied, the total current drain is only 3 mA.

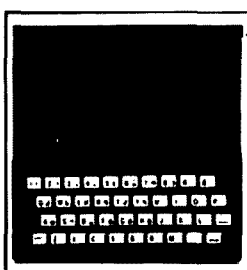
This field strength meter need not be limited to LF use only. **Table 1** shows the L1 and C1 values for HF operation and broadband operation.

operation

At first try, the added sensitivity was a blessing. I could easily make field strength measurements at distances that were impossible with the simple diode detector barefoot meter. At 200 feet from my LF antenna, testing on 186.5 kHz, I could easily get 30 percent scale readings.

I then began checking my 1750-meter antenna system. When the antenna was dry, I noted a 1-dB increase in field strength over readings taken when it was wet. I ran a 600-volt "Megger" test on a dry day and also on a rainy day, noting 10,000 Megohms on the dry day and 3 Megohms on the rainy day. In the past, under similar conditions, I couldn't detect this variation in field strength because of unusable readings. Noticing the substantial difference in field strength, I cut several tree branches that were touching my antenna and found that just doing this increased the field strength by another 2 dB. The field strength meter proved to be quite a useful tool.

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The most significant conversions appear in the section of the main program that reads back data files (**lines 1130-1190**) and in the file saving program shown in **fig. 2**. All the commands to the printer required modification, especially with the addition of **lines 5000-6220** to provide decimal alignment of the numerical results. The data tables from the original article must be typed in using the program shown in **fig. 2**.

reference

1. David G. Hopkins, VK4ZF, "Computer-aided Design of Long VHF Yagi Antennas," *ham radio*, May, 1986, page 28.

By Olin K. McDaniel, Jr., W4PFZ, 1327 Pinckney Avenue, Florence, South Carolina 29501

fig. 1. W4PFZ's revision of VK4ZF's Yagi design program runs on Commodore 64.

```
READY.

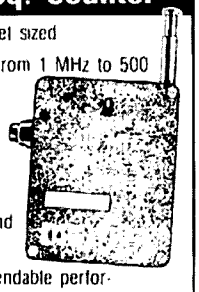
5 REM ********************************
10 REM       YAGI DESIGN PROGRAM
15 REM             BY
20 REM    DAVID G. HOPKINS (VK4ZF)
25 REM      #4 HANDSWORTH ST
30 REM          CAPALABA
35 REM        QLD.   4157
40 REM         AUSTRALIA
45 REM  ------------------------------
50 REM    MODIFIED FOR COMMODORE 64
55 REM             BY
60 REM     OLIN K. MCDANIEL (W4PFZ)
65 REM  ------------------------------
70 REM  PROGRAM BASED ON WORK DONE BY
75 REM       GUNTER HOCK (DL6WU)
80 REM        AND PUBLISHED IN
85 REM      'VHF COMMUNICATIONS'
90 REM  REQUIRES THE USE OF A PRINTER
100 REM
110 REM ********************************
120 REM
160 PRINT"{CLR}":GOSUB 1820
170 PRINT"{CLR}":DIM SP(40),DS(40),TS(40),LE(45),LE$(45),DS$(45):Q$="3333"
180 REM-------------LOAD ELEMENT SPACING DATA-------------
190 DATA.240,.075,.180,.215,.250,.280,.300,.315,.330,.345
192 DATA.360,.375,.385,.390,.395,.400,.400,.400,.400,.400
194 DATA.400,.400,.400,.400,.400,.400,.400,.400,.400,.400
196 DATA.400,.400,.400,.400,.400,.400,.400,.400,.400,.400
200 FOR X=1TO 40
210 READ SP(X)
220 NEXT
230 REM-------LOAD REFLECTOR MULTIPLIER----------
235 DIM RE(16)
240 DATA.4905,.4900,.4885,.4875,.4865,.4855,.4845,.4835
245 DATA.4825,.4820,.4810,.4785,.4770,.4765,.4750,.4740
260 FOR X=1TO16:READ RE(X):NEXT
270 REM-------LOAD RADIATOR MULTIPLIER-----------
280 DIM DR(16)
290 DATA.4675,.4665,.4640,.4620,.4601,.4585,.4575,.4550
295 DATA.4530,.4515,.4500,.4460,.4435,.4430,.4400,.4385
300 FOR X=1TO16:READ DR(X):NEXT
310 REM-------SELECT ELEMENT MATERIAL SIZE--------
315 DIM EL(16)
320 DATA.003,.0035,.0042,.005,.0056,.0063,.007,.0078
325 DATA.0088,.01,.01145,.0131,.015,.0165,.0182,.02
330 FOR X=1TO16:READ EL(X):NEXT
340 REM-------INPUT DESIRED PARAMETERS-----------
350 PRINT"{CLR}WHAT IS THE CENTER FREQUENCY OF THE     ANTENNA IN MHZ.":INPUT F
360 PRINT"{C/DN}HOW MANY ELEMENTS DOES THE ANTENNA     REQUIRE" :INPUT N
370 PRINT:IFN<9 OR N>40THENPRINT"{RVON}NUMBER OF ELEMENTS MUST BE BETWEEN 9 AND
40"
375 IFN<9 OR N>40 GOTO 360
380 PRINT"WHAT IS THE DIAMETER OF THE BOOM IN     MILLIMETERS":INPUT BD
390 PRINT:PRINT"ARE THE ELEMENTS TO BE INSULATED FROM   THE BOOM?    Y=YES, N=NO
 ";
400 GET A$:IF A$="" GOTO 400
410 IF A$="Y" THEN I=1:GOTO 450
420 IF A$="N" THEN I=0:GOTO 450
430 GOTO 400
440 REM---------DISPLAY SPECIFICATIONS-------------
450 PRINT"{CLR}":PRINT"SPECIFICATIONS FOR THE ANTENNA TO DESIGN":PRINT:PRINT
455 PRINT"1.  DESIGN FREQUENCY: ";F ;" MHZ"
460 PRINT"2.  DIAMETER OF BOOM: ";BD;" MM"
470 PRINT"3.  NUMBER OF ELEMENTS: ";N
480 IF I=1 THEN PRINT"4.  ELEMENTS ARE TO BE INSULATED FROM        THE BOOM"
484 IFI=0THEN PRINT"4.  ELEMENTS ARE TO BE ELECTRICALLY CON-NECTED TO THE BOOM"
490 IF(BD/1000)/(299.792/F) <.05 GOTO500
491 IF(BD/1000)/(295.792/F)>.05THENPRINT"{C/DN}{C/DN} BOOM DIAMETER TOO LARGE FO
R THIS",
492 PRINT"FREQUENCY":PRINT"{C/DN}{C/DN}":PRINT"ENTER A SMALLER DIAMETER":INPUT B
D:GOTO440
500 REM----------CHECK SPECIFICATIONS--------------
510 PRINT"{C/DN}{C/DN}{C/DN}":PRINT"ARE ALL THE ENTRIES CORRECT, Y=YES, N=NO";
520 GET A$:IF A$=""GOTO 520
530 IF A$="Y" GOTO 650
540 IF A$="N" GOTO 560
550 GOTO 520
560 PRINT"WHAT IS THE NUMBER OF THE INCORRECT ENTRY":INPUT X
570 IF X<>4 GOTO 600
580 IF I=0 THEN I=1:GOTO 440
590 IF I=1 THEN I=0:GOTO 440
600 INPUT"WHAT IS THE CORRECT VALUE";CV
610 IF X=1 THEN F=CV
620 IF X=2 THEN BD=CV
630 IF X=3 THEN N=CV
640 GOTO 450
650 REM----------CALCULATE BOOM DIAMETER WAVELENGTHS----------
660 W=299.792/F
670 BW=BD/1000/W
680 BC=BD*(526.286*BW1.648831/100)
690 REM----------CALCULATE BOOM LENGTH------------------
700 TL=0
```

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```
720 FOR X=1TO N-1
730 DS(X)=SP(X)*W*1000
740 TL=TL+DS(X)
750 NEXT
760 PG=(TL/1000)/W
770 REM----------CALCULATE BEAM WIDTHS AND STACKING DISTANCES----------
780 BH=50.2709*PG+.484081:SH=W/(2*(SIN(BH/2*.0174533)))
790 BV=66.5112*PG+.617261:SV=W/(2*(SIN(BV/2*.0174533)))
795 PA=INT((7.8*(LOG(PG)/LOG(10))+9.2)*1000+.55)/1000
800 REM----------PRINT PRELIMINARY DATA----------
810 PRINT"(CLR)"
820 PRINT"          YAGI PRELIMINARY DATA":PRINT
830 PRINT"FREQUENCY                ";F;" MHZ."
840 PRINT"WAVELENGTH               ",INT(W*10000)/10000," METERS"
850 PRINT"# OF ELEMENTS            ";N
860 PRINT"DIAMETER OF BOOM         ";BD;" MILLIMETERS"
870 PRINT"ELECTRICAL BOOMLENGTH    ";INT(TL);" MILLIMETERS"
880 PRINT"BOOM WAVELENGTHS         ";PG
890 PRINT"MAXIMUM PRACTICAL GAIN   ";PA;" DBD"
900 PRINT"HORIZONTAL BEAM WIDTH    ";:PRINTINT(BH*100+.5)/100,:PRINT"DEGREES"
910 PRINT"VERTICAL BEAM WIDTH      ";:PRINTINT(BV*100+.5)/100,:PRINT"DEGREES"
920 PRINT"(C/DN)              STACKING DISTANCES"
930 PRINT"         HORIZONTAL:  ";:PRINTINT(SH*1000+.55)/1000,:PRINT" METERS"
940 PRINT"         VERTICAL     ";:PRINTINT(SV*1000+.55)/1000,:PRINT" METERS"
950 PRINT"(C/DN)(C/DN)";:PRINT"DO YOU WISH TO CONTINUE WITH THIS DESIGN      Y=YE
S, N=NO"
960 GET A$:IF A$="" GOTO 960
970 IF A$="Y" GOTO 1000
980 IF A$="N" GOTO 440
990 GOTO 960
1000 PRINT"(CLR)":NU=0
1010 PRINT:PRINT"YOU MAY USE ANY OF THE FOLLOWING TUBING SIZES FOR THE ELEMENTS"
1020 PRINT:PRINT"SELECT THE SIZE CLOSEST TO A COMMERCIAL TUBE SIZE"
1030 FOR X=1TO 16
1040 NU=NU+1
1050 PRINT"#  ";X;:PRINTTAB(15)INT(((EL(X)*W)*1000)*1000+.55)/1000;
1055 PRINT" MM":NEXT
1070 PRINT"(C/DN)(C/DN)(C/DN)":PRINT"ENTER THE # OF THE TUBING SIZE YOU WISH TO
USE":INPUT TS$
1075 TS=VAL(TS$)
1080 IF TS<NU OR TS=NU THEN 1100
1090 IF TS>NU THEN PRINT"(RVON) ERROR (RVOF)":PRINT"HIT ANY KEY TO CONTINUE"
1093 IF TS>NU THEN GET A$:IF A$="" THEN 1093
1096 GOTO 1000
1100 TT=EL(TS):TD(1)=TT*W*1000
1110 PRINT"(CLR)":PRINT"STAND BY - THIS WILL TAKE A FEW SECONDS"
1120 REM----------LOAD TUBING SIZE TABLES----------
1125 DIM A$(38)
1130 N$="CURVE":N2$=N$+TS$
1135 OPEN15,8,15
1140 OPEN2,8,2,"0:"+N2$+",S,R"
1150 FOR X=1TO 38
1160 INPUT#2,TS(X)
1180 NEXT
1190 CLOSE2:CLOSE15:CLOSE8
1200 REM----------CALCULATE ELEMENT LENGTHS----------
1210 FOR X=3 TO 42
1220 IF I=1 THEN LE(X)=W*TS(X-2)*1000
1225 IF I=0 THEN LE(X)=W*TS(X-2)*1000*BC
1230 NEXT
1240 LE(1)=W*RE(TS)*1000
1250 IF I=0 THEN LE(1)=LE(1)*BC
1260 LE(2)=W*DR(TS)*1000
1270 IF I=0 THEN LE(2)=LE(2)*BC
1280 PRINT"(CLR)"
1290 PRINT"PRESS ANY KEY WHEN PRINTER IS READY"
1300 GET A$:IF A$="" GOTO 1300
1304 OPEN4,4:CMD4
1306 PRINT#4
1310 PRINT#4,"                YAGI DESIGN DETAILS:       "
1320 PRINT#4,"-----------------------------------------------------------------
1330 PRINT#4,"(C/DN)(C/DN)"
1335 GOSUB 5000
1340 PRINT#4,"DESIGN FREQUENCY:     ";SPC(37-AA);F$;" MHZ.";
1350 PRINT#4,"WAVELENGTH:           ";SPC(38-BB);W$;" METERS";
1360 PRINT#4,"NUMBER OF ELEMENTS:   ";SPC(36-NN);NO$;
1370 PRINT#4,"DIAMETER OF BOOM:     ";SPC(36-CC);BD$;" MM";
1380 PRINT#4,"DIAMETER OF ELEMENTS ";SPC(37-DD);TD$;" MM";
1390 IF I=1 THEN PRINT#4,"  'ELEMENTS ARE INSULATED FROM THE BOOM'"
1391 IF I=0 GOTO 1394
1392 GOTO 1400
1394 PRINT#4,"   'ELEMENTS ARE ELECTRICALLY CONNECTED TO THE BOOM'"
1400 PRINT#4,"ELECTRICAL BOOM LENGTH:     ";SPC(28-EE);TL$;" MM";
1410 PRINT#4,"BOOM WAVELENGTHS:           ";SPC(32-FG);PG$;
1420 PRINT#4,"MAXIMUM PRACTICAL GAIN:       ";SPC(30-FF);PA;" DBD";
1430 PRINT#4,"HORIZONTAL BEAM WIDTH:        ";SPC(29-GG);BH$;" DEGREES";
1440 PRINT#4,"VERTICAL BEAM WIDTH:          ";SPC(29-HH);BV$;" DEGREES";
1450 PRINT#4,"HORIZONTAL STACKING DISTANCE:";SPC(30-II);SH$;" METERS";
1460 PRINT#4,"VERTICAL STACKING DISTANCE:   ";SPC(30-JJ);SV$;" METERS";
1470 PRINT#4,"(C/DN)(C/DN)"
1480 PRINT#4,"            ELEMENT LENGTH, MM.     DISTANCE FROM REFLECTOR, MM."
1490 PRINT#4,"-----------------------------------------------------------------
---"
1500 PRINT#4,"(C/DN)"
1510 PRINT#4,"REFLECTOR",SPC(15-KK);LE$(1);SPC(29);"0";
1520 PRINT#4,"DRIVEN ELEMENT";SPC(10-LL);LE$(2);SPC(37-LL-LZ);SP$(1);
1530 PS=SP(1)*W*1000
1540 FOR X=3 TO N
1545 PS=PS+SP(X-1)*W*1000
1550 GOSUB 6000
1560 PRINT#4,"DIRECTOR #";X-2;SPC(13-MM-MN);LE$(X);SPC(37-MM-NN);PS$
```

```
1570 NEXT
1580 PRINT#4
1590 PRINT#4,"    ELEMENT SPACING IN MILLIMETERS    "
1600 PRINT#4,"--------------------------------------"
1610 PRINT#4
1615 GOSUB 6100
1620 PRINT#4,"REFLECTOR TO DRIVEN ";SPC(17-DD);DS$(1):
1630 PRINT#4,"DRIVEN TO DIRECTOR 1";SPC(17-PP);DS$(2):
1640 FOR X=3 TO N-1
1645 GOSUB 6200
1650 PRINT#4,"DIR";X-2;SPC(4-MN);"    TO DIR ";X-1;SPC(17-QQ-MD);DS$(X):
1660 NEXT
1670 PRINT#4:PRINT#4:PRINT#4,"           NOTES"
1680 PRINT#4:PRINT#4,"1. THE DIMENSIONS ARE FROM CENTER TO CENTER IN ALL CASES."
1690 PRINT#4,"      FOR EXAMPLE THIS MEANS THE BOOM MUST BE CUT LONGER THAN THAT"
1700 PRINT#4,"      GIVEN TO BE ABLE TO MOUNT THE ELEMENTS."
1710 PRINT#4,"2. IF YOU WANT WIDE BANDWIDTH, USE A FOLDED DIPOLE AS THE DRIVEN E
LE"
1720 PRINT#4,"3. THE DRIVEN ELEMENT DIMENSION IS THE LENGTH OVERALL."
1730 PRINT#4,"4. YOU MUST WORK TO AN ACCURACY OF <1MM AT FREQUENCIES ABOVE 400 M
HZ."
1740 PRINT#4,"5. ACCURACY BELOW 400 MHZ SHOULD BE WITHIN 1.5MM."
1750 PRINT#4,"6. ELEMENT MOUNTING MUST BE BETTER THAN .5MM OF THE ELEMENT CENTER
."
1760 PRINT"{CLR}":FOR X=1 TO 8:PRINT#4:NEXT
1770 END
1810 GOTO 1780
1820 PRINT"{CLR}":PRINT:PRINT"     THIS PROGRAM WILL DESIGN"
1825 PRINT"  YAGI ANTENNAS WITH ANY NUMBER"
1830 PRINT"    OF ELEMENTS FROM 9 TO 40."
1840 PRINT"{C/DN}{C/DN}{C/DN}"
1850 PRINT"  THE PROGRAM IS BASED ON ARTICLES"
1855 PRINT"   BY GUNTER HOCK AND PUBLISHED"
1860 PRINT"     IN 'VHF. COMMUNICATIONS'."
1870 PRINT"{C/DN}{C/DN}{C/DN}"
1880 PRINT"    YOU WILL REQUIRE A PRINTER"
1885 PRINT"     TO OBTAIN USABLE RESULTS."
1890 PRINT"{C/DN}{C/DN}    {RVON} PRESS ANY KEY TO CONTINUE {RVOF}"
1900 GET A$:IF A$=""GOTO 1900
1910 PRINT"{CLR}"
1920 GOTO 170
1930 END
5000 F=INT(F*100+.5)/100
5005 F$=STR$(INT(F))+"."+RIGHT$(STR$(F*100),2):AA=LEN(F$)
5010 W$=STR$(INT(W*1000)/1000):BB=LEN(W$)
5015 NO$=STR$(INT(N))+"."+RIGHT$(STR$(N*10),1):NN=LEN(NO$)
5017 BD=INT(BD*10+.5)/10
5020 BD$=STR$(INT(BD))+"."+RIGHT$(STR$(BD*10),1):CC=LEN(BD$)
5025 TD(1)=INT(TD(1)*100+.5)/100
5030 TD$=STR$(INT(TD(1)))+"."+RIGHT$(STR$(TD(1)*100),2):DD=LEN(TD$)
5040 TL$=STR$(INT(TL)):EE=LEN(TL$)
5042 PG=INT(PG*1000+.5)/1000
5045 PG$=STR$(INT(PG))+"."+RIGHT$(STR$(PG*1000),3):FG=LEN(PG$)
5050 PA$=STR$(PA):FF=LEN(PA$)
5055 BH=INT(BH*100+.5)/100
5060 BH$=STR$(INT(BH))+"."+RIGHT$(STR$(BH*100),2):GG=LEN(BH$)
5065 BV=INT(BV*100+.5)/100
5067 BV$=STR$(INT(BV))+"."+RIGHT$(STR$(BV*100),2):HH=LEN(BV$)
5070 SH=INT(SH*1000+.5)/1000
5080 SH$=STR$(INT(SH))+"."+RIGHT$(STR$(SH*1000),3):II=LEN(SH$)
5085 SV=INT(SV*1000+.5)/1000
5090 SV$=STR$(INT(SV))+"."+RIGHT$(STR$(SV*1000),3):JJ=LEN(SV$)
5095 LE(1)=INT(LE(1)*10+.5)/10
5100 LE$(1)=STR$(INT(LE(1)))+"."+RIGHT$(STR$(LE(1)*10),1):KK=LEN(LE$(1))
5105 LE(2)=INT(LE(2)*10+.5)/10
5110 LE$(2)=STR$(INT(LE(2)))+"."+RIGHT$(STR$(LE(2)*10),1):LL=LEN(LE$(2))
5120 SP$(1)=STR$(INT(SP(1)*W*1000))+"."+RIGHT$(STR$(SP(1)*W*1000*10),1)
5130 L2=LEN(SP$(1))
5200 RETURN
6000 LE(X)=INT(LE(X)*10+.5)/10
6010 LE$(X)=STR$(INT(LE(X)))+"."+RIGHT$(STR$(LE(X)*10),1):MM=LEN(LE$(X))
6013 EL$=STR$(INT(X-2)):MN=LEN(EL$)
6015 PS=INT(PS*10+.5)/10
6020 PS$=STR$(INT(PS))+"."+RIGHT$(STR$(PS*10),1):NN=LEN(PS$)
6030 RETURN
6100 DS(1)=INT(DS(1)*10+.5)/10
6105 DS$(1)=STR$(INT(DS(1)))+"."+RIGHT$(STR$(DS(1)*10),1):DD=LEN(DS$(1))
6110 DS(2)=INT(DS(2)*10+.5)/10
6115 DS$(2)=STR$(INT(DS(2)))+"."+RIGHT$(STR$(DS(2)*10),1):PP=LEN(DS$(2))
6120 RETURN
6200 DS(X)=INT(DS(X)*10+.5)/10
6205 DS$(X)=STR$(INT(DS(X)))+"."+RIGHT$(STR$(DS(X)*10),1):QQ=LEN(DS$(X))
6210 EL$=STR$(INT(X-2)):MN=LEN(EL$):EB$=STR$(INT(X-1)):MD=LEN(EB$)
6220 RETURN

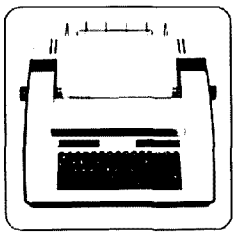
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(continued on page 102)

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VHF/UHF WORLD

Joe Reisert W1JR

RF connectors: part 1

Though the subject of RF connectors has been mentioned here many times, time and scheduling have not permitted a detailed discussion of this seemingly simple and straightforward subject.

I often hear Amateurs make all kinds of wild claims, or see them distribute misinformation on the subject. These stories typically involve power handling, impedance matching, and insertion loss. Also, many Amateurs seem unaware of the many types of connectors — besides the so-called "standard" or preferred types — that are available. These same Amateurs would probably also wonder why so many types are required.

This all reminds me of the story circulated in the early 1950s, just after the transistor was invented. The general notion was that all possible applications could be accommodated by the development of just a few types of transistors! Today there are over 10,000 types of numbered transistors, and new ones are becoming available weekly!

While there aren't nearly as many series of RF connectors available as of transistors, there surely must be almost as many types of distinctly different RF connectors and adapters available. Therefore, this month's column will serve as sort of a primer, in which we'll try to sift through the major series of RF connectors. Next month's column will expand on the various tradeoffs, especially in the area of applications. Tables will help put the whole selection of RF connectors in perspective and, I hope, help make selection easier in the future.

overview

RF connectors are designed to join or separate two components or units — such as an antenna and a transceiver — with relative ease. They're also used to gain access to a specific unit.

The proper choice of an RF connector type is particularly important when considering frequency of operation, passing high level RF power, or connecting to a low-noise preamplifier. It's also necessary if you're using a modular approach. Proper connectors permit convenient access to a unit so that adjustments can be made without disassembling or disturbing the circuit under test.

The actual choice of an RF connector depends on many things: application, availability, relative cost, electrical performance, mechanical durability, and environmental conditions. The choice of an optimum RF connector for a specific application, therefore, isn't always possible. Trade-offs are often necessary.

Furthermore, connectors are available in many forms. They may be threaded, bayonet, snap-on, or push-on. They may also be used on standard coaxial line, **Triax**™ (cable with two separately isolated braids for maximum shielding effectiveness), or **Twinax**™ (for connecting to 95-ohm balance line such as RG-22). The connector may be male, female, or neither.

An RF connector may be used either to terminate a transmission line, an in-series or between-series adapter, a panel mount, or feedthrough. What's the method used to attach to the connector to the line? Is it a solder connection, screw-on, push-on, or crimp type? Is the center pin free-floating or captive? Or is it fabricated from the actual transmission line center conductor itself, such as with RG-17 and UT-141 semi-rigid coax.

Basically speaking, RF connectors can be separated into three main families; standard, miniature, and subminiature. Standard connectors are for cables that are larger than 0.25 inches in diameter. Miniature connectors fit cable measuring between 0.10 and 0.25 inches. Subminiature connectors are sized for cables measuring less than 0.10 inches in diameter.

RF connector series

Table 1 is a list of most of the major series in common use, along with some technical data. More data will be provided in next month's column.

The **UHF** connector was the forerunner of most of the modern RF connectors. Developed in the mid-1930s, it was inexpensive and easy to assemble and use. It quickly became an industry standard. The most common types used by Amateurs are the PL-259 plug and the SO-239 chassis mount.

Unfortunately the UHF connector series doesn't have a constant impedance and is therefore usually limited to 500 MHz and below; in fact, I wouldn't recommend using UHF connectors above 150 MHz. What's more, it's not weatherproofed and therefore can't be recommended for outdoor use.

Connector development was spurred on by radar and VHF communications gear designed during WW II. Later progress resulted from the development of more demanding applications, especially at the higher microwave and millimeter-wave regions.

The **type N** connector — supposedly named for its inventor, Paul Neill of Bell Labs — was one of the first developed for both the VHF and UHF frequencies. Its most significant contribution to the state of the art was the addition of the separate outer contact, which wipes against the female body jack.

Table 1. This table contains some of the more pertinent parameters of the major series of RF connectors in common use.

Type	Outer diameter typical (inches)	Impedance Ohms	Thread size (inches)	Maximum Frequency
APC-3.5	0.312	50	1/4-36	34 GHz
APC-7	0.880	50	11/64-24	18 GHz
BNC	0.568	50 or 75	NA	4 GHz
C	0.750	50	NA	10 GHz
EIA	1 to 7	50	Note 1	Note 1
F	0.500	75	3/8-32	500 MHz
GR	0.812	50	NA	4 GHz
HN	0.875	Note 2	3/4-20	10 GHz
JCM	0.312	50	1/4-36	4 GHz
LC	1.500	Note 2	1 1/4-18	1 GHz
LT	1.500	50	1 1/4-18	1 GHz
MHV	0.562	Note 2	NA	50 MHz
N	0.825	50 or 70	5/8-24	11 GHz
RCA Phono	0.375	Note 2	NA	30 MHz
SC	0.812	50	11/64-24	11 GZHz
SM	0.312	50	1/4-32	200 MHz
SMA	0.312	50	1/4-36	12.4 GHz
SMB	0.250	50 or 75	NA	4 GHz
SMC	0.250	50,75,95	10-32	10 GHz
SSM	0.250	50	10-36	26 GHz
TM	0.425	50	5/16-32	10 GHz
TNC	0.625	50	7/16-28	10 GHz
Triax	1.000	Note 2	7/8-20	10 GHz
Twinax	.875	78 to 125	several	200 MHz
UHF	.750	Note 2	5/8-32	300 MHz

Note 1. Depends on diameter.
Note 2. May be used at 50- to 75-ohm impedances if matching is not a problem.

The N connector has a constant impedance and is usable to at least 11 GHz. Its gasket seals were later improved along with the braid clamp, making it an excellent choice for outdoor use. Amateurs usually prefer the UG-21 and UG-58 plug and jack, respectively. *One caution:* N connectors are sometimes made for 70-ohm impedances. To accomplish this, the diameter of the center pin is very small. Inserting a 50-ohm male connector into a female 70 ohm type will break the receptacle pin. Also, if a 70-ohm male "N" is inserted into a 50-ohm female N connector, there may be no electrical contact. I've seen these 70-ohm **N** connectors at flea markets, so beware. If you're not sure, compare the pin diameters with a known 50-ohm connector.

The **type C** connector, developed by Carl Concelman, is similar to the N connector. It was the first connector to use the bayonet-lock mating system so that it could be easily connected or disconnected. Though it's not as popular as the N connector, it's usable to at least 10 GHz.

Usable to 4 GHz, the **BNC** connector — quite popular with Amateurs, especially in the VHF and UHF region — has a bayonet-lock. Its name is said to derive from its developers, Neill and Concelman; hence its name, BNC, for (B)ayonet, (N)eill, and (C)oncelman [1]. The UG-88 plug and the UG-290 chassis jack are the most common BNC connectors seen in Amateur equipment. BNC connectors are sometimes available as 75-ohm connectors; the cautions about "N" type connectors apply to BNC connectors as well.

The **TNC**, an improved or "threaded" version of the BNC connector, has a more positive contact on the body and hence is recommended over the BNC connector, especially above 500 MHz. It's rated up to at least 10 GHz; a precision version is rated to 12.4 GHz.

Widely distributed by the Omni Spectra Manufacturing Corporation, which called it an "OSM" connector, the **SMA** connector is now manufactured by many companies. This connector, defined as a miniature type, is popular on UHF and above, especially where low-loss, constant impedance, and small size are required.

The SMA connector was primarily developed to mate with 0.141-inch semi-rigid metal jacketed cable [2]. In the early 1970s, the E.F. Johnson Company introduced the **JCM** connector, a lower-frequency version of the SMA connector which is specified to 4 GHz and also mates with RG-174 miniature flexible cable. This low-cost connector is very popular in UHF applications.

SSM subminiature connectors were designed later. Primarily suited for 0.085 semi-rigid coax, these work up to 26 GHz. Improved SSM connectors will work up to 40 GHz!

The **SMB** and **SMC** connector series

is primarily used in video and IF equipment, where small-diameter flexible cabling is used. The SMB is threaded; the SMC is a push-on type. Both are relatively inexpensive; SMC is the more common of the two, especially on military equipment.

The **APC 3.5** is the ultimate in precision. Resonance-free through 34 GHz, it will mate directly with the SMA connector series.

The **APC-7** is an expensive precision connector used primarily for 0.250-inch semi-rigid coax. It's a hermaphrodite, which making it an excellent connector for instrumentation.

An **EIA** connector is very unusual in that it doesn't have a threaded or bayonet connection. Instead it consists of body and moveable flange that's bolted to the mating connector. It's primarily used on air lines, where low-loss, high-power 7/8 to 6 1/8-inch lines such as on Heliax (TM) are used.

The inexpensive "**F**" connector is popular on CATV installations, especially for connecting units together with RG-59-type 75-ohm cable. Usually a crimp-type, it often uses the center wire of the cable for the center pin. Unfortunately, it's not rugged, weather-proofed, or very usable above about 300 MHz.

General Radio Corporation designed the **GR** connector primarily for instrumentation — hence its initials. A hermaphrodite push-on type connector most commonly seen as the G-874 or GR900 types, it's especially good on patch panels and where equipment must be frequently connected or disconnected.

The **HN** connector is basically a larger-diameter offshoot of the N connector with a higher voltage rating. It's primarily used with large-diameter (7/16 to 7/8-inch) cables when breakdown voltages up to 5000 volts are required.

LC and **LT** connectors are large and very similar. They're usually seen where high voltages or high power are required. Often used with RG 17-type coax, both of these connector series use the mating cable with its dielectric as the center portion of the connector.

The **MHV** connector may at first appear to be a BNC connector, but that's where the similarities end. The MHV is primarily designed to handle high voltages (up to 5000 volts) with a quick connect/disconnect capability. This feature is especially useful for high-voltage power supplies for tube type amplifiers, as described in reference 3. The MHV is seldom used for passing RF.

Normally I wouldn't even mention **RCA** phono connectors, except that I often see them on older commercial gear through 500 MHz! They're definitely not recommended for RF use above 30 MHz because they have no specific impedance properties. They're best suited for audio and control circuits where shielded cables are required.

The **SC** connector, a screw-on version of the C connector, is often used in applications requiring higher voltage than the C connector can handle and where environmental conditions are more severe.

The **SM** connector isn't too common. It's not weatherproofed and is recommended only for IF work. It may be used at different impedances if match isn't important.

A subminiature version of the SMA connector, the **SSM** is primarily used where the smallest possible size and highest possible frequency are required.

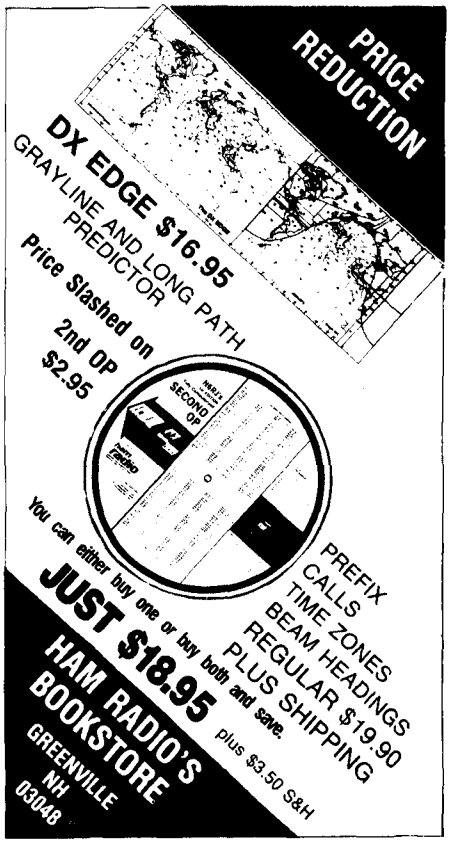
The **TM** series isn't often seen. Basically, it's a 2/3 size version of the TNC connector sometimes seen on IF connections.

The **Triax** and **Twinax** connector series are used when balanced lines are required. The Triax type allows a cable with two insulated shields to be used. The Twinax connector provides for superior shielding. Some Twinax connectors are polarized, while others are not. They're popular on video transmission and computer installations.

adapters

So far we've been discussing only coaxial connectors such as plugs, jacks, and chassis types. But other RF connectors function as adapters, both within and between series adapters.

There may be as many different RF adapter types as cable connectors. The



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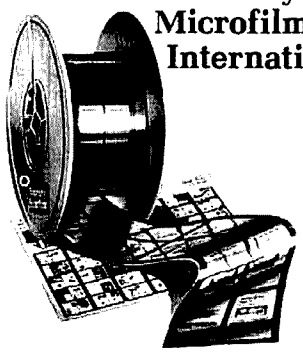
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MASTER LOG
Over three years of development went into this program. It creates a file of 2100 individual records with up to 13 different entries per record. Master Log can do a search and select based upon time, frequency, mode or any of the other variable parameters. It keeps track of DXCC and WAS status, prints QSL labels and can search its whole file in less than 5 seconds! Complete documentation is included to help you learn and use this truely state-of-the-art logging program.
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most common ones are probably the female-to-female or male-to-male barrells for connecting cables together in the same series. Between-series adapters are also common. Right-angle and "T" types are popular since they allow flexibility when you want to hook similar connectors together.

It's particularly important to choose the proper adapter. I often see installations with two or even three adapters placed in series even though there's a single adapter available that will make the connection. The loss through each adapter is about two to three times that of a normal connector.

Your best bet is to choose a single adapter with the desired transition between series. Flea markets seem to have just about every imaginable type. If you're in doubt, bring one of your connectors along and check the fit.

Be particularly careful when using adapters outdoors. The more adapters used, the more likely it is that moisture will seep into one of the connections.

next month

Next month's column will go further into the mechanical and electrical aspects of connector design and use. Tips on proper use will also be given. This information should aid in your choice for the optimum connector for your application.

references

1. Allen Nemetz, "A Designer's Guide to RF Connector Selection," *rf design*, September/October, 1980, page 18.
2. Joe Reisert, W1JR, "VHF/UHF World: Transmission Lines," *ham radio*, October, 1985, page 83.
3. Joe Reisert, W1JR, "VHF/UHF Word: High Power Amplifiers — Part 2," *ham radio*, February, 1985, page 38.

Important VHF/UHF Events:

Sept. 6-7	*International Region 1 VHF Contest*
Sept. 12	*EME perigee*
Sept. 13-14	*ARRL VHF QSO Party*
Sept. 26-27	*First ARRL 10 GHz Cumulative Contest, first weekend*
Oct. 4-5	*Mid-Atlantic States VHF Conference, Warminster, PA (contact WA2OMY)*
Oct. 4-5	*International Region 1 UHF/SHF Contest*
Oct. 7	*EME perigee*
Oct. 10-11	*First ARRL 10 GHz Cumulative Contest, second weekend.*
Oct. 20	*Predicted peak of the Orionids Meteor Shower at 1645 UTC*
Oct. 25/26	*International EME contest, first weekend*

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low-cost spectrum analyzer
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Although laboratory-grade spectrum analyzers cost $4500 or more, you can build a spectrum analyzer offering many features of its costlier cousins for about $50. How can such amazing capabilities be had at such incredibly low cost?

- *Through the use of a commercially mass-produced varactor tuned TV tuner that covers the VHF low, VHF high, and UHF TV bands.* It will tune down through the 2-meter and 6-meter Amateur bands without modification, with better than 1-microvolt sensitivity. By using a crystal-controlled converter and a narrower IF filter, any one of the HF Amateur bands could be viewed as well, by simply upconverting to a TV channel.
- *Through the use of consumer-grade integrated circuits in the oscillator/mixer, dual ceramic filter, IF amplifiers/detector, and audio amplifier (offering audio as well as scope output it is really a spectrum monitor).*
- *Through the use of your own oscilloscope.* Just about any scope may be used; I used a 1951 Heathkit Model OL-1 with its original cathode ray tube.

spectrum analyzer applications

Spectrum analyzers allow the user to observe in real time an adjustable/variable bandwidth of radio frequencies. One of the earliest spectrum analyzers for the ham bands was the "Panadaptor," manufactured by Hallicrafters in the early 1950s. I used one to check for F2 propagation 6-meter band openings in the mid-1950s and credit the panadaptor with my earning the IARU 6-meter WAC award (phone) issued by the ARRL.

The band of frequencies swept by the spectrum analyzer described in this article may be varied from zero up to about 38 MHz on VHF low TV = 50 MHz – 88 MHz, zero up to about 85 MHz on VHF high TV = 135 MHz – 220 MHz, and zero up to about 300 MHz on the UHF TV = 500 MHz – 800 MHz. When the sweep width is set at zero, each of these bands of frequencies may be manually tuned just as in a single-frequency receiver. Both wideband FM and narrowband FM signals, and surprisingly, even amplitude modulated signals are detected quite well by the FM IF amp/detector IC and amplified by the audio IC.

Figure 1 is a block diagram of this spectrum analyzer using a Sanyo varactor tuned TV tuner. A ten-turn, 10k pot with +35 VDC across it is used to adjust the center frequency of the tuner. A low varactor bias yields low frequency and a high varactor bias provides higher frequency on the TV band to which the tuner is set. The sawtooth sweep voltage that is capacitively coupled into the tuner's varactors is the horizontal sweep voltage from the oscilloscope. If your scope doesn't have the horizontal sweep output (the Heathkit OL-1 does not), just mount an RCA phono jack on the front of the scope and bring out the horizontal sweep from the scope's horizontal multivibrator to this point. The 100 kilohm load across the horizontal sweep output should have little or no effect on the scope's operation.

Besides the TV tuner and scope, the rest of the circuit consists of only three integrated circuits. The second mixer/oscillator chip is a Siemens SO42P. The ceramic filter is a Murata SFJ two-section filter at 10.7 MHz. The combination IF amp/detector/AGC amp is a National LM-3089N, and the 1/2-watt audio amp is a National LM-386N.

construction, testing, and alignment

Figure 2 is a schematic of the analyzer. A printed circuit board is available from WA2PZO; I recommend using this and WA2PZO parts kits. (Although the pots, S-meter, and speaker aren't furnished, most are available from Radio Shack.) **Figure 2A** shows the component layout on the WA2PZO printed circuit board (foil side down). Alternatively, you could use perfboard and point-to-point wiring.

Figure 3 is a schematic diagram of the interconnections between the Sanyo tuner, the three-IC printed circuit board, and the scope. Once everything is connected as shown in **fig. 3**, alignment can begin. (It should be easy, since there are only four adjustments,

By Robert M. Richardson, W4UCH, 22 North Lake Drive, Chautauqua, New York 14722

the ferrite cores in L1, L2, L3, and L4.)

With all the parts on hand — and assuming you're using the PCB — assembly time is at most an hour or two. I tuned up my unit using only a grid dip meter as a signal source in about 20 minutes.

Construction of the PCB proceeds as follows:

1. Install the jumper at the lower edge of U3 as illustrated in **fig. 2A**.
2. Install the 11 resistors as shown in **fig. 2A**.
3. Install the 16 capacitors as indicated in **fig. 2A**. Since C6 and C7 are electrolytics, be sure to observe the polarity indicated in **fig. 2A**.
4. Install the Murata ceramic filter, FL1. Because it's symmetrical, it can be installed in either direction.
5. I recommend carefully installing sockets for U1 (16-pin DIP), U2 (8-pin DIP), and U3 (14-pin DIP).
6. Install L1 through L5, carefully bending the pins slightly so that they fit easily into the PCB's pre-drilled

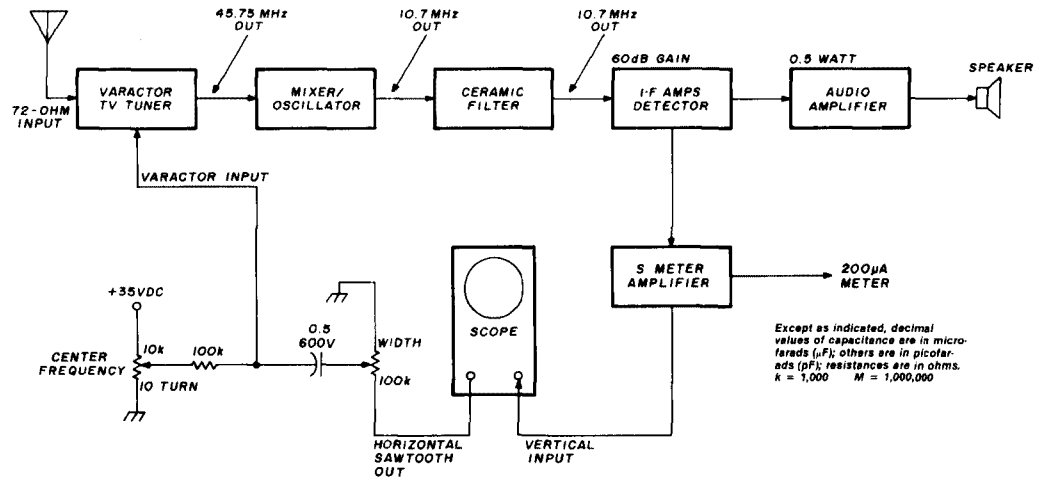


fig. 1. Block diagram of WA2PZO spectrum analyzer.

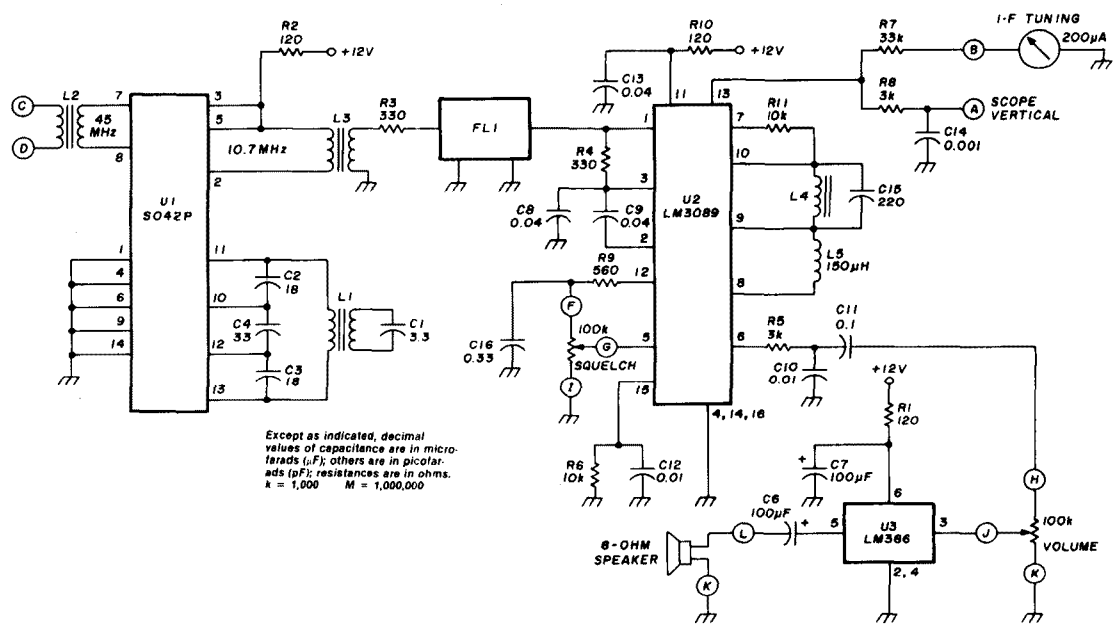


fig. 2. Schematic of WA2PZO spectrum analyzer.

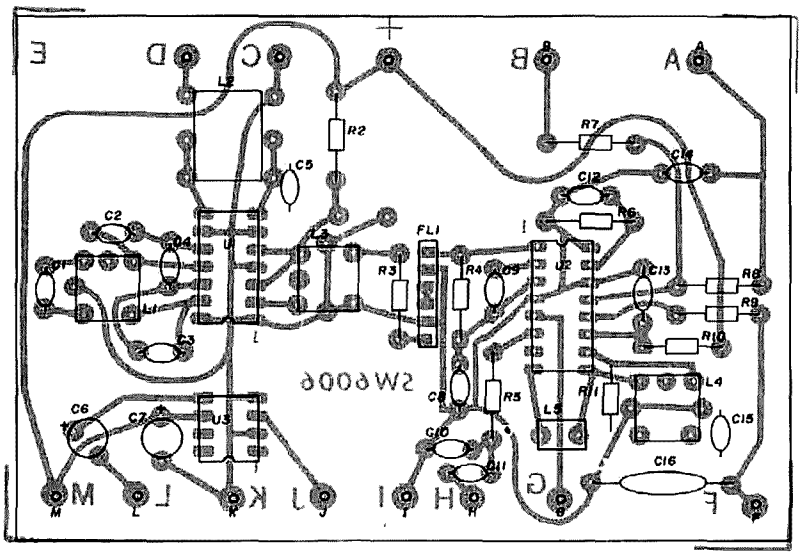


fig. 2A. Component layout on the WA2PZO printed circuit board (foil side down).

holes. Because L5, the 150 μH inductor, is symmetrical it can be installed in either direction.

7. Most important: lubricate the small ferrite cores of L1, L2, L3, and L4 with a drop or two of WD-40 *before* attempting adjustment. These cores are extremely brittle; use a plastic or nylon tuning tool matched to the width of the ferrite slots (using a metal one will guarantee their destruction).
8. If you're not using a 200-μA S-meter, install a jumper from 'A' to ground.

Testing and alignment proceed as follows:

1. Connect the TV tuner's VHF input to your 2-meter antenna.
2. Set the VHF/UHF switch to the VHF position.
3. Set the Hi/Lo switch to *Hi*.
4. Make sure the top of the ferrite core in L1 is exactly even with the top of its shielded housing.
5. Set the squelch control to zero and adjust the volume control for a slight hiss from the speaker with the gain control at maximum resistance (10 kilohms).
6. Set the width control (sawtooth sweep from scope) to zero.
7. Connect a voltmeter to pin 10 of the TV tuner (varactor input). Use the +30-volt voltmeter range (adjust the center frequency 10k-turn pot so that varactor bias is below +30 VDC).
8. Turn the center frequency pot slowly down (reducing varactor bias voltage) until a local TV video carrier is both heard on the speaker (buzz) and seen on the scope. If no TV stations are nearby, use a signal generator with modulated output or a grid dip meter.

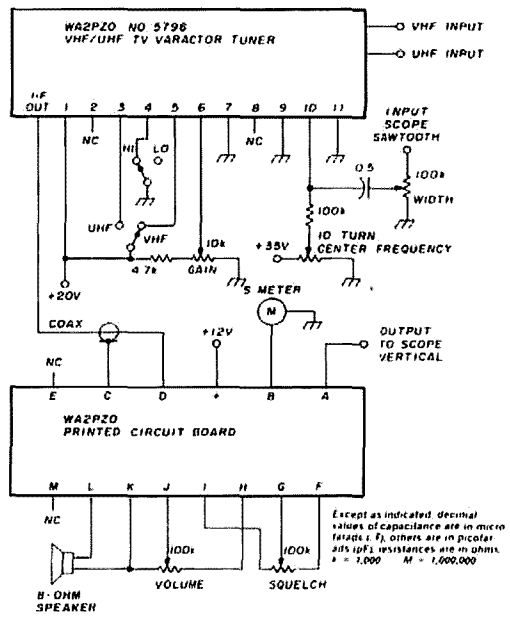


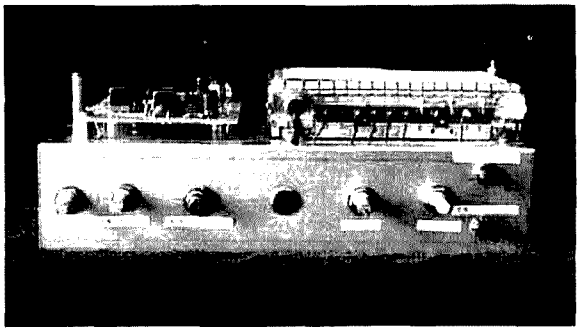
fig. 3. Schematic showing interconnections between tuner, printed circuit board, and oscilloscope. Note: +35, +20, and +12 VDc sources are regulated supplies.

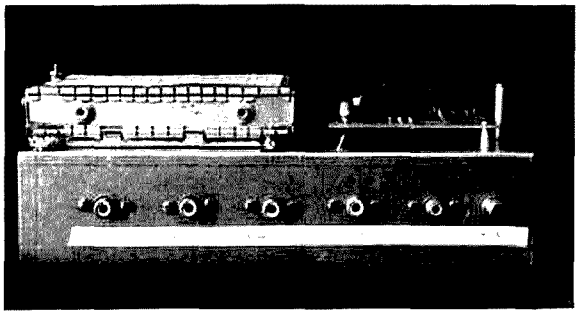
9. Peak the ferrite slugs in L2 and L3 for maximum signal output on the scope.
10. Continue tuning the center frequency pot until a TV audio signal is found, then adjust L4 for maximum

audio output and minimum background noise from the speaker.

11. Though each varactor tuned TV tuner will have somewhat different voltage versus frequency response, see **fig. 10** to see the response we obtained using a Sanyo varactor tuned TV tuner.

Figure 4 is a photo of our finished spectrum analyzer. The tuner and PCB are mounted on top of a 2 x 5 x 9-inch (5 x 13 x 23 cm) aluminum chassis. The PCB is mounted on 1/2-inch (1.3 cm) threaded standoffs. Left to right, the five pots are: volume, squelch, center frequency, sweep width, and tuner gain. The two mini-toggle switches on the right of **fig. 4** are VHF-UHF (top) and VHF LOW—VHF HIGH (bottom). On the rear of the chassis are six RCA phono jacks: +35 VDC for tuner center frequency; +20 VDC for tuner; +12 VDC for oscillator/mixer, second IF amps, U3 audio amp; horizontal sweep from scope; vertical output to scope; and audio output to the 8-ohm speaker.





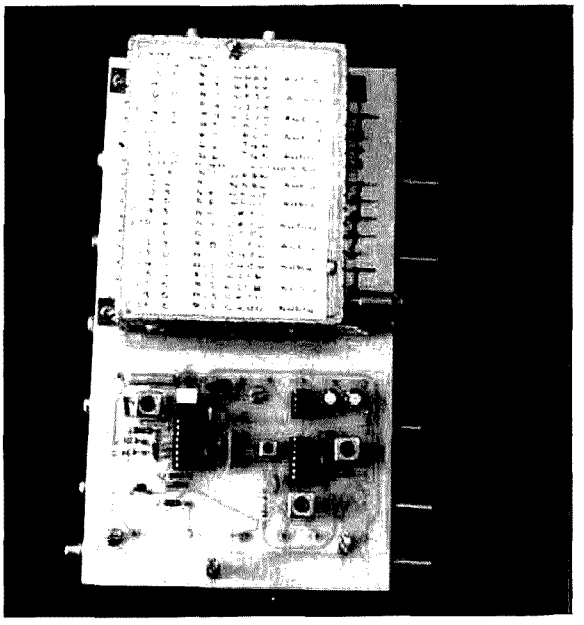


fig. 4. Completed spectrum analyzer: (A) front, (B) rear, (C) top.

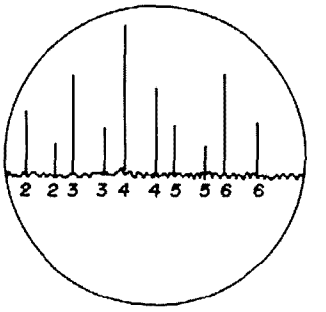


fig. 5. VHF low TV band with horizontal sweep amplitude set to maximum.

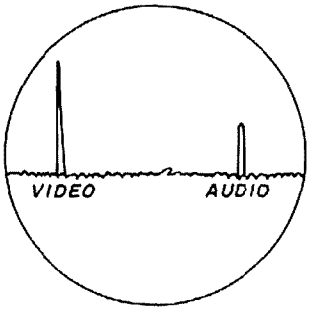


fig. 6. 8-MHz horizontal sweep width, with channel 2 video signal at left and channel 2 audio at right.

operation

Now the fun really begins. For antennas, I used my two 23-element Cushcraft 2-meter "Boomers." (No, I don't bounce signals off the moon with them, but I can work into 2-meter repeaters in Toronto, some 125 miles away. My QTH is on the south shore of Chautauqua Lake, some 65 miles southwest of Buffalo; because of this distance, and the presence of a range of hills about 300 feet (90 meters) high between Buffalo and Chautauqua, TV signals aren't par-

ticularly strong. Nevertheless, all the TV channels on both the VHF low TV and VHF high TV bands were displayed on the first try. Both the vestigial sideband video signal and its accompanying FM audio signal were clearly displayed for each channel.

Figures 5 through **10** are sketches of the oscilloscope display in the noted modes of operation. (Oscilloscope cameras cost about $300; I chose not to add one to my $39.95 Heathkit scope.)

Figure 5 shows the VHF low TV band with the horizontal sweep amplitude set to maximum. Each TV channel pair displayed (video and audio) is noted. The height of each signal is proportional to signal strength; some Canadian TV stations really pack a wallop even though they're 125 miles away.

Figure 6 illustrates about 8 MHz horizontal sweep width (left side of CRT to right side of CRT) with the video signal of Channel 2 on the left and the audio from Channel 2 on the right side.

Figure 7 displays the sweep width reduced to about 2 or 3 MHz with the center frequency set to Channel 2's video carrier. Note the blanking and vertical sync pulse riding on top of the carrier.

Figure 8 is the Channel 2 video carrier with the horizontal sweep set to zero. The blanking pulse with the vertical sync pulse on top is on the left side. The eight squiggles to the right of the vertical sync pulse on top of the blanking pulse are the color burst; all the hazy, wavy signals to the right are the video information. The top of the vertical sync pulse represents 100 percent modulation and the bottom ot the video information represents the white level of video at about 15 percent modulation.

Although some scope photos or sketches show 100 percent modulation at the bottom, I prefer it at the top. If you insist on having it at the bottom, simply turn the figure upside down and view it in a mirror.

Figure 9 illustrates the 2-meter band with horizontal sweep representing about 3 MHz. The left side of the CRT is at 145 MHz and the right side of the CRT is at 148 MHz. Spread between 146 and 148 MHz, we can see about six 2-meter repeaters located in the Buffalo and Toronto areas.

Figure 10 is a plot of varactor tuning voltage versus frequency on my Sanyo TV tuner. The 10-turn, 10-ohm pot used for setting the varactor voltage is an absolute "must" for fine tuning.

By reducing the sweep width to zero and single-signal tuning across each band, I was able to copy the audio on the VHF TV low band from the following stations: WGRZ-TV (Channel 2, Buffalo) WPSX-TV (Channel 3, Rochester), WIVB-TV (Channel 4, Buffalo), CBLT-TV (Channel 5, Canada), and CTGN-TV, (Channel 6, Canada). On the VHF TV high band we copied audio from air-to-ground and air traffic control stations; 2-meter repeaters in western New York and the Toronto area; commercial FM pagers; Toronto Coast Guard marine weather on 161.775 MHz; the Erie, Pennsylvania, weather bureau on 162.40 MHz; the Buffalo, New York, weather bureau on 162.55 MHz; WKBW-TV (Channel 7, Buffalo); WROC-TV (Channel 8, Rochester); CFTO-TV (Channel 9, Canada); CFPL-TV (Channel 10, Canada); CHCH-TV (Channel 11, Canada); WICU-TV, (Channel 12, Erie); and CKCO-TV (Channel 13, Canada). On the UHF TV

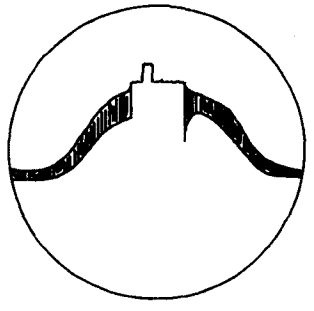


fig. 7. Sweep width reduced to about 2 or 3 MHz with center frequency set to channel 2's video carrier.

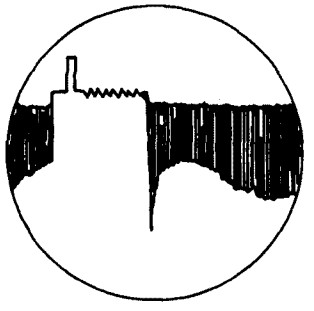


fig. 8. Channel 2 video carrier with horizontal sweep set to zero.

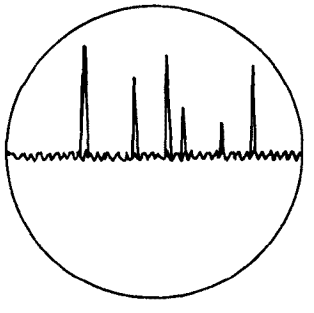


fig. 9. 2-meter band with horizontal sweep representing about 3 MHz.

Fig. 10. Varactor voltage versus frequency (will vary from tuner to tuner).

VHF low band

voltage	frequency MHz
0.5	50.00 6-meter band
2.0	55.25 channel 2 video
3.5	59.75 channel 2 audio
4.0	61.25 channel 3 video
6.0	65.75 channel 3 audio
7.0	67.25 channel 4 video
9.5	71.75 channel 4 audio
13.5	77.55 channel 5 video
17.1	81.75 channel 5 audio
20.1	83.25 channel 6 video
34.8	87.75 channel 6 audio

VHF high band

0.5	135.00 air to ground
1.0	144.00 2-meter band
2.0	148.20 commercial paging
4.0	162.55 Buffalo weather
6.5	175.25 channel 7 video
7.6	179.75 channel 7 audio
7.9	181.25 channel 8 video
9.0	185.75 channel 8 audio
9.3	187.25 channel 9 video
10.6	191.75 channel 9 audio
11.0	193.25 channel 10 video
12.3	197.75 channel 10 audio
13.0	199.25 channel 11 video
14.7	203.75 channel 11 audio
15.2	205.25 channel 12 video
17.5	209.75 channel 12 audio
19.0	211.25 channel 13 video
25.5	215.75 channel 13 audio

band we copied about six TV stations from Channel 17 through Channel 26, including a French language Canadian TV station. The Sanyo TV tuner is one super little box with excellent sensitivity — i.e., much better than 1 microvolt. All these stations were copied using only 2-meter antennas.

other uses

Once you become accustomed to using it, the spectrum analyzer is probably the greatest trouble-shooting aid since Volta invented the voltmeter. With the proper probes, attenuators, and converters it can be used as an RF voltmeter; as a signal tracer in transmitters and receivers; and for transmitter alignment, harmonic measurement, deviation measurement, oscillator injection measurement, IF alignment, and spurious radiation measurement, to name but a few of its applications. If you want to dig deeper, Cushman Electronics (2450 North First Street, San Jose, California 95131) publishes a neat little book entitled *Using The Spectrum Monitor* Priced at $7.25 (postpaid), it's easily understood by the average Radio Amateur.

acknowledgement

I wish I could claim authorship for this unique design, but credit goes entirely to my friend, Murray Barlowe, WA2PZO, who not only created the design but has generously made a partial kit of parts, TV tuner, and printed circuit board available to Amateurs at virtually his cost.

For more information send an SASE to Murray Barlowe, WA2PZO, P.O. Box 393, Bethpage, New York 11714. Once you try it, you'll never understand how you got along without it. **ham radio**

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PRACTICALLY SPEAKING ...

Joe Carr K4IPV

substituting transistors part 1: using resources at hand

Every Amateur who spends at least some time at the workbench repairing or building electronic equipment will eventually need a transistor that's not on hand — and perhaps not even available. In some cases, the type number will be found in one of the standard transistor replacement catalogs. In other cases, well . . . you're on your own.

Although the subject of transistor substitution is one that's been talked about to the point of exhaustion among hams, serious problems continue to reappear. The tips given in this two-part series, while most appropriate to the types of transistors normally used in Amateur Radio equipment, are also applicable to a variety of other situations as well.

The main premise is that we are *servicing* Amateur Radio equipment that once worked properly and then failed. While much of what is discussed is also applicable to construction projects, construction project debugging is something of an arcane art and is thus not suitable for general, too-broad guidelines. I can recall several projects over the years that depended for proper operation on selected parameters of specific transistors, and wouldn't even work with all otherwise-working versions of the same "2N" number devices.

There are even cases on record in which only those devices made by certain manufacturers will work properly in the circuit. (Ancient history note: those of us who go back to vacuum tube days recall a very costly ham receiver that would retain its "frequency meter" dial calibration only when RCA tubes were used for the local oscillator; there are transistor equivalents to that situation.) As a result, we must limit our consideration to repair of working — and, one would hope, properly designed — equipment.

exact replacements

The easiest way to obtain a replacement solid-state device that will install easily and operate correctly is to order it from the original equipment manufacturer or an authorized distributor. As we're all painfully aware, however, this is not always either possible or practical.

industry-standard type numbers

If the defective transistor has a standard "2N" type number, then you just obtain a replacement device with the same number, without regard to brand name. Unfortunately, some "Original Equipment Manufacturer" (OEM) transistors aren't marked with these standard numbers. They often have a house code number that's meaningless to anyone except the manufacturer. Sometimes the house number is created because the transistor is specially selected from others for the same "2N" series, so only a similarly tested device will work properly in the circuit. In some cases, the house number is used because it suits the manufacturer's inventory control system; in other cases, manufacturers simply want to ensure replacement parts business.

crossover guidelines

Crossover guides, which would seem to be a nearly perfect source of replacement numbers, should be used whenever possible. But there are gremlins that can pop up unexpectedly. Theoretically, the cross-matching has been done in advance by the use of an "infallible" computer. When we follow those recommendations, however, we sometimes find that suggested replacements have insufficient power or voltage ratings, too narrow a bandwidth, a different physical shape that would cause mounting or space problems, or different mounting dimensions that would require modification of the chassis. Many of these discrepancies occur because the crossovers are compiled from printed lists that sometimes contain errors. It's an open secret that the recommended substitutes are seldom tried in any kind of equipment or circuit, so it's best to test the reasonableness of any selection by looking at the crossover device's specifications and comparing them with what you know about the circuit and its requirements.

During my years in the electronic service business, it was my policy to return, along with a note of explanation, any crossover transistors that either didn't work properly or would require major reworking of the chassis or rewiring of the circuit. If everyone did this, manufacturers might take the hint. The economic impact of a service shop's annual semiconductor purchases makes it easy for them to obtain refunds on bad crossovers; unfortunately, Amateurs rarely have such clout.

Another problem has nothing to do with electrical specifications, but rather with proper identification. In some cases it's relatively easy to guess the required transistor type. But what if two manufacturers have each accidentally assigned the same designation to two

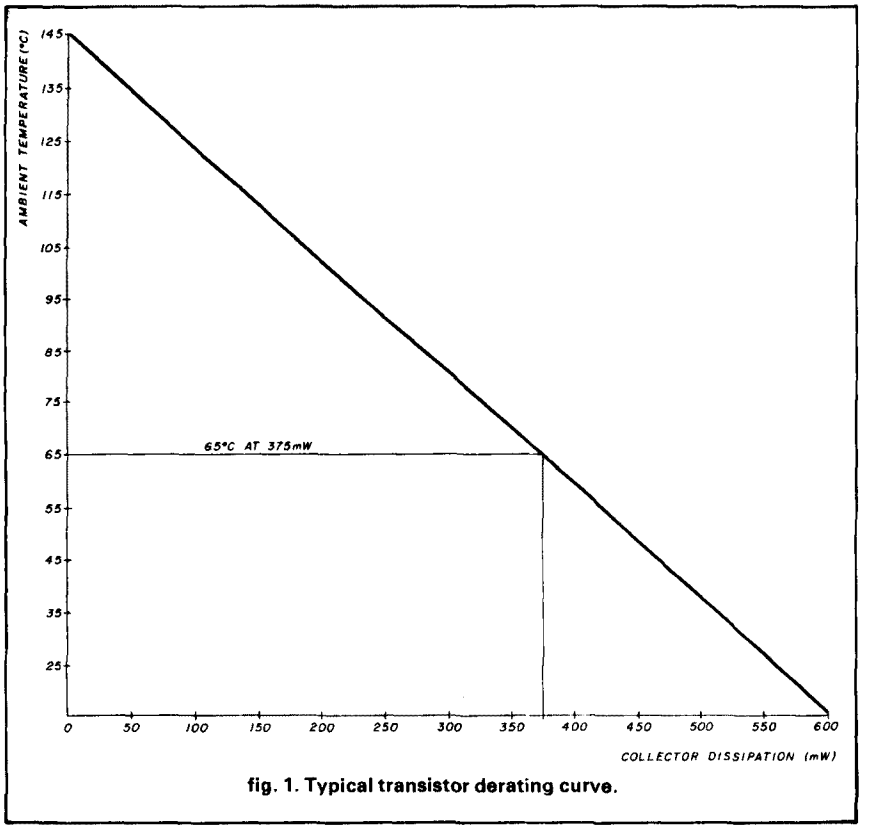


fig. 1. Typical transistor derating curve.

completely dissimilar devices? It's not likely that a crossover guide will solve all such problems, though some do accommodate such ambiguities.

I remember one case years ago where I needed a replacement for a Delco Electronics DS-25. Now, the DS-25 has been around for about 20 years as an RF amplifier, IF amplifier, and converter replacement in Delco-General Motors car radios. The DS-25 germanium transistor was packaged in a "smaller-than-TO-5" case. Unfortunately, a small hi-fi manufacturer also used the DS-25 designation for a medium-power PNP germanium power transistor in a TO-3 case. One crossover guide I consulted at that time listed the TO-3 type without noting that it wasn't the Delco part number, even though it was listed among the various Delco "DS-Series" type numbers.

Remember the old rule from high school math: *Things equal to the same thing are — you hope — equal to each other.* Or, if A = B, and B = C, then A = C. We can use this observation to make crossover selections. Furthermore, we can use this technique in at least two additional ways. First, we can look up the device needed to find the replacement type number. For example, suppose a 2N5xxx is found in the Zotch Electronics Crossover Guide as a "ZE-234." We can look for other "2N" series devices also equal to ZE-234 and use one of those.

This method is especially useful when crossing house numbers to 2N numbers, which is our second way to use the "A = C" theory. Suppose that your Wombat Thunderbolt VI transceiver uses a transistor with the part number "8501234" Well, the Zotch guide calls it a ZE-234. By looking over the "2N" series columns in the *Zotch Guide To Replacement Things*, you find that a 2N5xxx is also equal to a ZE-234. Chances are good that the Wombat engineers selected the 2N5xxx and then relabeled it "8501234." It may not be the exact transistor, but it's a fair bet that it will work unless the 8501234 is a specially selected 2N5xxx. (There are no guar-

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P/N		Rating	Each	Match Pr.
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MRF421	Q	100W	22.50	51.00
MRF422*		150W	38.00	82.00
MRF426,/A*		25W	18.00	42.00
MRF433		12.5W	12.00	30.00
MRF449,/A	Q	30W	12.50	30.00
MRF450,/A	Q	50W	14.00	31.00
MRF453,/A	Q	60W	15.00	35.00
MRF454,/A	Q	80W	15.00	34.00
MRF455,/A	Q	60W	12.00	28.00
MRF458		80W	20.00	46.00
MRF475		12W	3.00	9.00
MRF476		3W	2.75	8.00
MRF477		40W	11.00	25.00
MRF479		15W	10.00	23.00
MRF485*		15W	6.00	15.00
MRF492	Q	90W	16.75	37.50
SRF2072	Q	65W	13.00	30.00
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	Rating	MHz	Net Ea.	Match Pr.
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MRF224	40W	136-174	13.50	32.00
MRF237	4W	136-174	3.00	—
MRF238	30W	136-174	13.00	30.00
MRF239	30W	136-174	15.00	35.00
MRF240	40W	136-174	18.00	41.00
MRF245	80W	136-174	28.00	65.00
MRF247	75W	136-174	27.00	63.00
MRF260	5W	136-174	7.00	—
MRF261	10W	136-174	9.00	—
MRF262	15W	136-174	9.00	—
MRF264	30W	136-174	13.00	—
MRF607	1.75W	136-174	3.00	—
MRF641	15W	407-512	22.00	49.00
MRF644	25W	407-512	24.00	54.00
MRF646	40W	407-512	26.50	59.00
MRF648	60W	407-512	33.00	69.00
SD1441	150W	136-174	74.50	170.00
SD1477	100W	136-174	32.50	78.00
2N3866*	1W	30-200	1.25	—
2N4427	1W	136-174	1.25	—
2N5591	25W	136-174	13.50	34.00
2N6080	4W	136-174	7.75	—
2N6081	15W	136-174	9.00	—
2N6082	25W	136-174	10.50	—
2N6083	30W	136-174	11.50	24.00
2N6084	40W	136-174	13.00	31.00

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MRF137	24.00	S10-12	13.50
MRF138	35.00	2SC1075	25.00
MRF140	89.50	2SC1307	5.00
MRF150	89.50	2SC1946A	12.00
MRF172	62.00	2SC1969	3.00
MRF174	80.00	2SC2221	10.00
2N1522	7.95	2SC2269	20.00
2N4048	7.20	2SC2289	22.00
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antees, however. You still have to check specs and make an educated guess).

mobile problems

Up to this point, the problems of obtaining suitable replacements for transistors in mobile rigs are identical to those of home-based rigs. But there's one special problem that's more serious in mobile applications: environmental heat.

One car manufacturer became concerned about the excessive number of failures in his first all solid-state car radio models (circa 1962) and decided to investigate the possibility of passenger cabin heat as the culprit. The company asked its electronics plant employees to leave their car doors unlocked for one day. During that day of 90-degree weather, the engineers measured the temperatures inside many of the closed cars and were surprised to find that the average reading was 160 degrees F on the seat and 180 degrees behind the dashboard.

derating the specs

Published transistor power ratings are usually specified at room temperature, generally accepted to be 25 degrees C (77 degrees F). If transistors are used at higher temperatures, as in mobile applications, then the maximum collector dissipation (in watts) must be reduced to prevent extra failures which could occur even when all the electrical specifications are fulfilled.

A typical derating curve is show in **fig. 1.** Notice that a transistor having a collector dissipation of 550 mW at 25 degrees C can safely dissipate only 375 mW at 65 degrees (142 degrees F). This explains why a transistor that's operating below its maximum published wattage rating could be destroyed by using it in a hot car. Watch for such hazards, especially if you attempt to use "five-for-a-buck" bargain-basement replacements in which the collector dissipation rating was "optimistic."

beyond crossmatching

Often the numbers on a bad transistor seem meaningless. There's no way to crossmatch and you can't locate an OEM replacement from the equipment maker. The next step is to find a universal replacement from one of the many convenient sources. To do this, you have to become an electronic detective and find out the following things about the transistor:

- Is it a silicon or germanium type?
- Is it a PNP or NPN?
- What is the gain (alpha or beta)?
- What frequencies must it amplify?
- What are the collector power dissipation requirements?
- Are there any special mechanical mounting requirements?

Once you've answered these questions, you can make a satisfactory selection from most any brand of universal replacements.

silicon or germanium?

Silicon transistor junctions measure higher DC resistances than germanium junctions. In fact, silicon transistors usually read "open" on all measurements except with base/emitter and base/collector forward biasing polarity. If even one junction remains intact on a blown transistor, then you can tell which material it is by comparing readings with those of known types in the size and power category.

Forward bias voltages for all stages other than oscillators (and certain pulse circuits used in video equipment) should be 0.2 to 0.3 volts for germanium transistors and 0.5 to 0.7 volts for silicon transistors. Check the schematic, or another of the intact junctions (or a similarly numbered good transistor in the circuit) to see which voltage levels are found. The answer will tell you whether to look for Ge or Si transistor replacements.

PNP or NPN?

When the collector voltage is more positive (or less negative) than the emitter, then the transistor is an NPN type. If the collector is more negative (or less positive) than the emitter, then the transistor is a PNP type. Most schematics give these voltage readings. On the other hand, you can measure the collector/emitter voltages accurately enough for this purpose, right in the circuit in most cases.

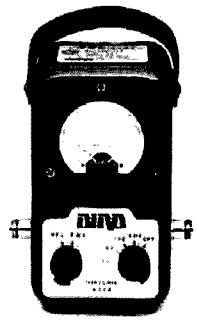
Even if only one junction of the defective transistor is intact, you still can determine the polarity by using the ohmmeter. If you obtain a normal low-resistance diode-type reading with the positive ohmmeter lead on the base and the negative lead on the collector or emitter, then it's an NPN type. If you must reverse the leads to obtain a low-resistance reading, the transistor is a PNP type. This measurement must be made with an old-fashioned VOM/VTVM or a modern digital meter with a "diode" or "high power" ohmmeter function.

next month

Next month we'll look at the frequency response of the selected transistor and certain mechanical considerations.

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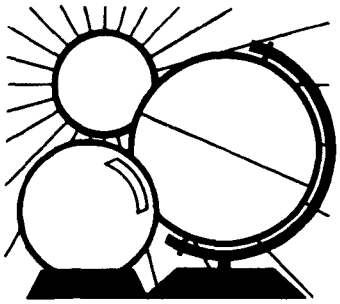
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DX FORECASTER

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equinox season DX

Even though this is a sunspot minimum year, and the sun's activity has decreased, an occasional energy burst (flare) or increase in the solar wind causes a geomagnetic field disturbance. We experienced these phenomena in February and May of this year, and it's likely that we'll have another in September or October. Whenever they occur, you can expect them to affect your DXing.

Geomagnetic disturbances, or storms, affect propagation — and DX — in three ways. First, particles from the sun entering the auroral zone at 50 to 70 degrees North and South latitudes come down into the ionospheric D and E regions, increasing signal absorption. This results in weak east-west path signals and few transpolar signals. Second, the particles form a reflective curtain along the equatorial side of the auroral zone (for those of us in North America, this is south), enhancing VHF auroral scatter propagation. Six-meter openings to Europe are one result of this phenomenon. Third, the F region of the ionosphere (for U.S. stations, this is south of the auroral zone) has a depleted area of electrons that form an electron density trough. The maximum usable frequency (MUF) for paths through this area decreases by 30 to 40 percent. (Tables of MUF statistics were presented in this column in January, 1986.)

However, still further south at ±20 degrees from the geomagnetic equator, an equivalent-size enhancement of the F region occurs, resulting in evening Trans-Equatorial (TE) openings during the equinox and winter seasons. These three effects vary in intensity and time on a short to long basis (seconds through hours), causing what we experience as fading and blackout. These effects continue to occur each night for two to three days before ionospheric equilibrium is re-established. The larger the geomagnetic storm (the higher the value of the K or A indicies), the closer to the equator these effects occur.

Just as the particle density and speed of the solar wind vary, so do the characteristics of the geomagnetic field and ionosphere. Ionospheric variation causes signal reflection focusing and defocusing, which simply means that the signals arriving at your QTH will vary in both strength and angle of arrival. Some directions and locations you haven't heard from in a long time may suddenly be workable.

last-minute forecast

The higher-level 27-day sunspot activity may push the maximum usable frequencies up during the first and last weeks of September, giving better 10-, 12-, and 15-meter DX. September marks the beginning of the return of transequatorial one-long-hop propagation for the winter season; during some evenings, it will probably be useful for DX. Its effect will be enhanced by an equinoxial increase in geomagnetic disturbances, which are more probable near the end of the second week and into the third.

The lower bands will experience less QRN caused by weather storm frontal thunderstorms passing through. But the geomagnetic disturbances will have greater effect on these bands; lower MUFs will occur on east, west, and north paths. Signal strength variability, QSB, is also associated with the disturbances. Listen carefully for new, unusual DX openings at these times.

A full moon will occur on September 26th and its perigee on the 12th. The autumnal equinox will be on the 23rd at 0759 UTC. No significant meteor showers are expected this month.

band-by-band summary

Six meters may have a few sporadic E openings around local noon, but don't count on them this last month of the season.

Ten, twelve, and fifteen meters should provide a few short-skip openings and many long-skip openings to most southern areas of the world, especially if there is any solar flux increase during the daylight hours this month. Some of these openings will result from transequatorial propagation, mainly during disturbed conditions.

Twenty, thirty, and forty meters will support propagation from east, west, and north areas of the world during the daytime and into evening hours almost every day. Distances to 2000 miles via long-skip or some short-skip Es to 1000 miles per hop are usual.

Thirty, forty, eighty, and one-sixty meters are all good for nighttime DX. The bands will be open in the east soon after sundown, swing toward the north and south about midnight, and end in the Pacific areas during the hour or so before dawn. The time-and-frequency stations in England and Hawaii make good band monitors. On some nights these bands will be as good as they are during the winter DX season; on others, QRN may be a problem. Distances will be a little shorter than those mentioned above.

WESTERN USA

GMT	PDT	N	NE	E	SE	S	SW	W	NW
0000	5:00	20	40	20	10	12	10	10	15
0100	6:00	20	40	20	12	12	10	10	20*
0200	7:00	20	40	20	12	15	10	10	20
0300	8:00	20	40	20	15	15	10	10	20
0400	9:00	20	40	20	20	20	10	10	20
0500	10:00	30	40	20	20	20	12	12	30
0000	11:00	40	40	20	20	30	12	12	30
0700	12:00	40	40	20	20	30	15	15	30
0800	1:00	40	40	20	20	30	20	20	30
0900	2:00	40	40	30	20	30	20	20	40
1000	3:00	40	40	30	30	30	20	20	40
1100	4:00	40	30	20	20	40	20	20	40
1200	5:00	40	20	15	20	40	20	20	40
1300	6:00	30	20	12	20	40	20	20	40
1400	7:00	30	20	10	15	40	30	20	40
1500	8:00	40	20	10	12	20	30*	20	40
1600	9:00	40	20	10	12	20	30	20	40
1700	10:00	40	20	10	10	20*	20	20	40
1600	11:00	40	20	10	10	15	20	30	30
1800	12:00	40	20	12*	10	15	15	20	20
2000	1:00	40	20	12	10	12	12	15	20
2100	2:00	40	20	15	10	12	12	12	20
2200	3:00	30	30	20*	10	12	12	12	20
2300	4:00	20	40	20	10	12	10	10	15
SEPTEMBER									
ASIA FAR EAST		20	40	20	10	12	10	10	15
EUROPE		20	40	20	12	12	10	10	20
S. AFRICA		20	40	20	12	15	10	10	20
S. AMERICA		20	40	20	15	15	10	10	20
ANTARCTICA		20	40	20	20	20	10	10	20
NEW ZEALAND		30	40	20	20	20	12	12	30
OCEANIA AUSTRALIA		40	40	20	20	30	12	12	30
JAPAN		40	40	20	20	30	15	15	30

MID USA

MDT	N	NE	E	SE	S	SW	W	NW	CDT
6:00	20	40	20	12	15	10	10	20	7:00
7:00	30	40	20	12	15	10	10	20	8:00
8:00	30	40	20	15	15	10	10	30	9:00
9:00	40	40	20	20	20	10	10	40	10:00
10:00	40	40	20	20	20	12	10	40	11:00
11:00	40	40	20	20	30	12	12	40	12:00
12:00	40	40	20	20	30	15	12	40	1:00
1:00	40	40	20	20	30	20	15	40	2:00
2:00	40	40	20	20	30	20	15	40	3:00
3:00	40	40	20	30	30	20	20	40	4:00
4:00	20	40	15	20	30	20	20	40	5:00
5:00	20	30	12	20	40	20	20	30	6:00
6:00	20	20	12*	40	20	20	20	30	7:00
7:00	20	20	10	15	40	20	20	40	8:00
8:00	20	20	10	12	40	30	20	40	9:00
9:00	40	20	10	12	30	30	20	40	10:00
10:00	40	20	10	10	20	30	20	40	11:00
11:00	40	20	10	10	20*	20	20	40	12:00
12:00	40	20	10	10	15	20	20	40	1:00
1:00	40	20	12*	10	15	15	20	20	2:00
2:00	40	20	12	10	15	12	15	20	3:00
3:00	40	20	15	10	15*	12	12	20	4:00
4:00	30	30	20*	10	15	12	12	20	5:00
5:00	20	40	20	10	15	10	10	20	6:00
ASIA FAR EAST	20	40	20	10	12	10	10	15	
EUROPE	20	40	20	12	12	10	10	20	
S. AFRICA	20	40	20	12	15	10	10	20	
S. AMERICA	20	40	20	15	15	10	10	30	
ANTARCTICA	20	40	20	20	20	10	10	40	
NEW ZEALAND	30	40	20	20	20	12	10	40	
OCEANIA AUSTRALIA	40	40	20	20	30	12	12	40	
JAPAN	40	40	20	20	30	15	12	40	

EASTERN USA

EDT	N	NE	E	SE	S	SW	W	NW
8:00	30*	40	20	12	15	10	10	20
9:00	30	40	20	12	15	10	10	20
10:00	40	40	20	15	20*	12	12	30
11:00	40	40	20	20	20	12	15	40
12:00	40	40	20	20	20	15	15	40
1:00	40	40	20	20	30	20	20	40
2:00	40	40	20	20	30	20	20	40
3:00	40	40	20	20	30	20	20	40
4:00	40	40	30*	20	30	20	20	40
5:00	20	20	20	30*	40*	20	20	40
6:00	20	20	15	20	40	20	20	30
7:00	20	20	12	20	40	30*	30	30
8:00	20	20	12*	15	40	20	20	40
9:00	20	20	10	12	40	20	20	40
10:00	20	20	10	12	20	20	20	40
11:00	40	20	10	12	20	20	20	40
12:00	40	20	10	10	20	20	20	40
1:00	40	20	10	10	20*	20	20	40
2:00	40	20	10	10	15	20*	20	40
3:00	40	20	12*	10	15	15	15	20
4:00	40	20	12	10	15	12	12	20
5:00	40	30	15	10	15*	12	12	20
6:00	40	40	20*	10	15	10	10	20
7:00	30	40	20	10	15	10	10	20
ASIA FAR EAST	30	40	20	10	15	10	10	20
EUROPE	30	40	20	12	15	10	10	20
S. AFRICA	40	40	20	15	20*	12	12	30
CARIBBEAN S. AMERICA	40	40	20	20	20	12	15	40
ANTARCTICA	40	40	20	20	20	15	15	40
NEW ZEALAND	40	40	20	20	30	20	20	40
OCEANIA AUSTRALIA	40	40	20	20	30	20	20	40
JAPAN	40	40	20	20	30	20	20	40

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.

*Look at next **higher** band for possible openings.

NEW products

RS 232-compatible computer interface units

Trio-Kenwood Communications has announced the release of RS 232-compatible computer interface units for the TS-440S, TS-940S, TS-711A, and TS-811A transceivers.

Two units are required to control the transceivers: the IF-232C level translator and the appropriate plug-in computer interface module. The TS-440S requires the IC-10 chip set; the TS-940S, the IF-10B; and the TS-711A or TS-811A, the IF-10A.

All digital functions on the transceivers – including VFO tuning, RIT/XIT, memory input and recall, and voice synthesizer activation – are controllable,. Programming is simple; one program should work with several rigs.

The suggested retail prices are IF-232C, $49.95; IF-10A/IF10-B, $41.95; IC-10, $22.95.

A simplified sample program will be available from Kenwood dealers. Write to Trio-Kenwood Communicatiions, 1111 West Walnut Street, Compton, California 90220, for information.

PCB kits from Kepro

A new line of Kepro pre-packaged materials and kits includes the following:

- a standard manual-resist etched circuit kit for producing basic PC boards;
- a photo-reversing kit for making line negatives from artwork prepared on transparent film or reversing negatives to positives;
- an immersion tin-plating kit to improve solderability by depositing 0.00001 inch of tin on the oxide-free copper surface of etched PCBs;
- photo-resist etched circuit kits, basic artwork or master photo layout kits, screen printing or nameplate kits.

In addition, Kepro also offers KeproClad,™ for the production of industrial quality, negative acting dry film photo-sensitized PCBs. KeproClad is available with foil on one or two sides, in sizes ranging from 4 x 6 through 7 x 12 inches, and is priced as low as $3.50.

For details, contact Kepro, Inc., 630 Axminster, Fenton, Missouri 63026.

Circle #307 on Reader Service Card.

New computer-based instruments from Heath

Models IC-4802 and ID-4850 oscilloscopes from Heath Company are designed to work with personal computers.

The IC-4802 Digital Oscilloscope is a sophisticated interface that attaches to an IBM PC compatible computer and is available in kit form or assembled. This interface turns an IBM PC compatible computer into a full-featured 50-MHz dual trace oscilloscope that allows full control of the scope from the keyboard of the computer, harnessing the computer's computational abilities. With the IC-4802, oscilloscope waveforms can be collected and stored on disk for later recall. The digitally stored waveforms may be printed out on the computer's printer.

The ID-4850 Digital Memory Oscilloscope is an interface that may be used with either a personal computer or a 5-MHz or greater bandwidth oscilloscope that has the ability to trigger from an external source and triggered sweep. Used with a computer, the ID-4850 provides a PC-compatible computer with 50-MHz oscilloscope capabilities and allows waveforms to be digitally stored for later recall. When using an oscilloscope, the ID-4850 upgrades it to a full-featured 50-MHz dual trace oscilloscope. The Digital Memory Oscilloscope is available in kit or assembled form.

More information on these and other products is available in Heathkit's free catalog. For a copy, contact Heath Company, Dept. 150-775, Benton Harbor, Michigan 49022.

Circle #309 on Reader Service Card.

personal frequency standard

Wenzel Associates, Inc., offers the new Counter-Mate personal frequency standard that provides stable 1-MHz and 10-MHz reference signals to improve the accuracy of counters and other instruments. A precision third-overtone 10-MHz crystal is mounted in a controlled oven in an installation that provides minimum aging and drift. Both outputs will drive TTL or 50 ohms with 5-ns rise and fall time square waves. The output impedance properly matches power splitters for generating isolated signals to operate several instruments. The price is $350.

Further information is available from Wenzel Associates, Inc., 11124 Jollyville Road, Austin, Texas 78759.

Circle #308 on Reader Service Card.

new Hamtronics catalog

Hamtronics, Inc. has announced publication of their new 40 page, two-color catalog, which features many new products, including several GaAs FET preamps, a five-function DTMF decoder/controller, a transmit/receive relay module, digital FSK equipment, and packet-radio VHF power amplifiers. Also included is a comprehensive listing of FCC type-accepted transmitters, receivers, and repeaters for commercial service. Hamtronics reports that because of recent high volumes of production, it has been able to reduce prices on many products.

To receive a copy by return first class mail, send $1 ($2 for overseas mailing) to Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535.

surge protector

Alpha Delta has announced availability of a new, improved version of its Transi-trap Electrical Surge Protector.

The new Transi-trap "Arc-plug" has been redesigned to meet government and industry protection standards for Electromagnetic Pulse (EMP) in accordance with the National Communications System report, NCS TIB 85-10. The "ARC-plug" has a DC clamping level of 350 volts to provide proper transmitter protection. The pulse clamping level (per NCS EMP test, 4,500 volts at 50 ohms) is 230 volts. The unit will respond in 80 to 100 nanoseconds and has a very low interelectrode capacitance of less than 1 pF.

The Transi-trap design offers low loss – typically 0.1 dB through 500 MHz for the R-T and 0.3 dB loss through 1 GHz for the units with N connectors.

For more information, contact Alpha Delta, P.O. Box 571, Centerville, Ohio 45459.

Circle #304 on Reader Service Card.

RF test equipment catalog

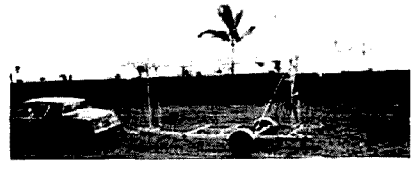
A new 60-page catalog of Thruline® directional wattmeters, coax load resistors and attenuators, calorimeters and RF components is available from Bird Electronic Corporation.

Included are such items as high-accuracy instruments using plug-in elements with 5000-to-1 power ranges, a frequency/power meter combination and relative field-strength devices, as well as more than 300 standard RF products.

This reference work of RF measurement instrumentation and components from 2 milliwatts to 250 kilowatts in the frequency range of 0.235 to 2300 MHz features triple indexing – by function, power level and model number – making it easy to use as both a desktop reference and specification tool.

Catalog GC-86 is available from Bird Electronic Corporation, Cleveland (Solon), Ohio 44139-2794.

Circle #312 on Reader Service Card.



trailer-mounted towers

Trailer-mounted communication towers, available from Aluma Tower Company, are well-suited for mobile testing, site selection for earth stations, civil defense, or other applications requiring a temporary communication tower. Towers up to 100 feet can be provided with either manual crank mechanisms or 12-V winch operation. Trailers are supplied with a 2-inch ball hitch, spring suspension, and tail lights for day/night service.

Contact Aluma Tower Company, Inc., 1643 Old Dixie Highway, Vero Beach, Florida 32961, for information.

Circle #303 on Reader Service Card.

new multimode xcvr from Yaesu

Yaesu U.S.A. has announced the release of the new FT-767GX, the world's first HF/VHF/UHF multimode transceiver. The FT-767GX comes factory-equipped for HF operation on the Amateur



bands plus general coverage on receive from 100 KHz to 29.99 MHz). Features include are an automatic antenna tuner for 160-10 meters, CW filter, electronic keyer, speech processor, digital wattmeter, IF shift, IF notch filter, CW audio peak filter, and a dual VFO tracking system for OSCAR Mode A or repeater operation. All popular operating modes are included: SSB, CW, AM, FM, and FSK. Optional modules extend coverage to 6 meters, 2 meters, and/or 70cm, and an optional CTCSS unit is available for tone-access repeater work.

The FT-767GX is compatible with Yaesu's CAT (Computer Aided Transceiver) external computer protocol, for remote control operation and enhanced operating flexibility. For teletype operation, the FT-767GX is rated at 100 watts output continuously for up to 30 minutes.

The introductory price of the FT-767GX is $1759.95, with the 6-meter and 2-meter modules priced at $169.95 each. Prices are subject to change due to the extreme volatility in international exchange rates.

For information, contact Yaesu, Inc., 17210 Edwards Road, Cerritos, California 90701.

Circle #305 on Reader Service Card.

DX beam heading chart

John Daley, KB6JGH, has announced the availability of his DX Beam Heading Chart. Each report is individually calculated by computer and packaged in an attractive 8-1/2 x 11-inch binder. Nine categories of data including callsign area, country, state, city, longitude, latitude, beam heading, Great Circle distance, nautical miles, and statute miles are provided for each of 540 DX locations. All data are are based on the user's exact station location.

Geared to both the ham and SWL, numerous listings are given for the USSR and China, making unknown site estimation easier.

The price of this chart is $9.95. It's available from John Daley, KB6JGH, P.O. Box 4794, San Jose, California 95150.

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ICOM IC2AT/12AT, IC02AT, IC-04AT, IC-A2/U16
Land Mobile HT Midland/Standard Wilson Mason Yaesu FTC 2203, FT 4703 Icom IC-M12 (Marine) M700 Tempo M-1

TS440S/AT, R-1000, R-2000, TS-040 S/AT
TM-201B, TR2600/3600A, TM2570A/50A/30A
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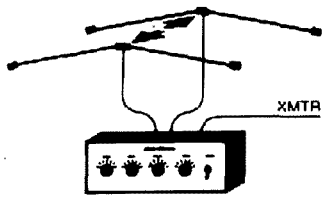
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(continued from page 75)

fig. 2. W4PFZ's "File Saver" program allows data files to be saved.

```
READY.

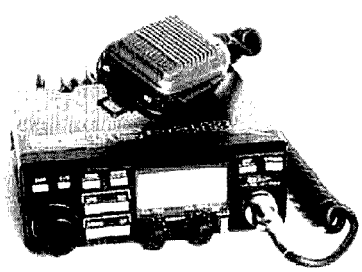

10 PRINT"{CLR}"
20 DIM A(39):Z$="CURVE"
30 PRINT"{C/DN}WHAT IS THE NUMBER OF THE CURVE YOU WISH TO ENTER "
35 INPUT C$
40 F$=Z$+C$
50 PRINT"ENTER THE VALUES FOR THE CURVE"
60 FOR X=1 TO 38
70 PRINT"#"; X;:INPUT A(X)
80 NEXT
90 PRINT"{CLR}"
100 PRINT"THESE ARE THE VALUES FOR CURVE";C$
110 PRINT"{C/DN}{C/DN}"
130 FOR X=1TO 38
140 PRINT"#";X;A(X)
150 NEXT
160 PRINT
170 PRINT"ARE ALL THE VALUES CORRECT?                    Y=YES  N=NO"
180 GET A$:IF A$="" THEN 180
190 IF A$="Y" GOTO 250
195 IF A$="N" GOTO 210
200 GOTO 170
210 PRINT"WHICH # IS INCORRECT":INPUT D
220 PRINT"WHAT IS THE CORRECT VALUE FOR :- #";D
230 INPUT A(D)
240 GOTO 100
250 PRINT"{CLR}"
280 PRINT"SAVING TO DISK"
265 OPEN15,8,15
270 OPEN2,8,2,"0:"+Z$+C$+",S,W"
280 FOR X=1 TO 38:PRINT#2,A(X)
300 NEXT
310 CLOSE 2:CLOSE15:CLOSE8
320 PRINT"{C/DN}{C/DN}{RVON}  SAVING COMPLETE  {RVOF}"
330 END

READY.
```




IC-48A 440-MHz mobile

ICOM has announced the release of the new IC-48A 440-MHz compact mobile. The IC-48A offers the same features as the new IC-28A and IC-28H, with 440-450MHz frequency coverage. Features include compact size (5-1/4 x 5 1/2 x 2 inches); a large LCD readout, with an automatic dimmer; 21 memory channels; scanning; plus an internal speaker and an HM-12 mic. With only 11 front panel controls, the unit is easy to operate.



Options include the IC HM14 DTMF mic, PS-45 13.8-volt power supply, UT-29 tone squelch unit, SP-10 external speaker, HM-16 speaker mic and HS-15/HS-15B flexible boom mic, and PTT switchbox.

For information, contact ICOM America, Inc., 2380 116th Avenue NE, Bellevue, Washington 98004.

Circle #313 on Reader Service Card.

Satellites Today — 2nd edition

Universal Electronics has announced the release of the enlarged second edition of Frank Baylin's popular book, *Satellites Today*, containing the latest satellite information.

Satellites Today, which can be understood by a non-technical reader, reviews satellite history and technology. Topics include uplinking, footprints, programming, home satellilte TV systems, and more.

Retailing for $12.95, *Satellites Today* can be ordered from the publisher, Universal Electronics, Inc., 4555 Groves Road, Suite 13, Columbus, Ohio 43232.

Circle #301 on Reader Service Card.

Edek "R" series enclosures

The Edek "R" series enclosures are distinctive, ruggedly-built enclosures suitable for the hobbyist, engineer, or anyone needing an attractive cabinet.

Dark walnut-stained natural wood ends, contrasted with a bright aluminum chassis, are featured. If a different color scheme is desired, the ends and chassis may be painted, or the aluminum chassis may be sanded for a brushed look. The ends may be grooved with a hand saw for PCB mounts; heavy components may be fastened with wood screws. The galvanized steel bottom is easily soldered or drilled. Non-marring rubber feet conceal hidden screws.

Enclosures are available in an unlimited number of sizes, with no minimums and no tooling charge. Delivery is from stock for the most popular sizes. Many variations are possible on the "R" series. A sample is available for $3.00 plus $2.00 shipping. A new slope-front enclosure ("S" series) will be available soon.

For information, contact Energy Engineering, Route 4, Fayetteville, Arkansas 72701.

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9-volt Rechargeable NiCads

Plainview Electronics announces the introduction of its 9-volt rechargeable NiCad "transistor" battery. Unique in a rechargeable transistor battery is its 8.4-volt nominal voltage, 110 mAh, and its low self-discharge characteristic, which will maintain 50 per cent capacity even after a year in storage at room temperature.

For information, contact Plainview Electronics, 28 Cain Drive, Plainview, New York 11803.

Circle #311 on Reader Service Card.

callbook supplement

The new combined *Callbook Supplement* includes all the changes in both the North American and International Callbooks for the six months since the publication of the regular Callbooks. Published once a year on June 1st, it lists *thousands of new licensees, address changes,* and "then and now" call changes from countries around the world.

Unlike previous updates, this new Supplement is available through regular Callbook dealers and is priced lower than previous Supplements.

For more information, contact Ham Radio's Bookstore or the publisher, Radio Amateur Callbook Inc., 925 Sherwood Drive, Lake Bluff, Illinois 60044.

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0508	50-54	170	1	-	-
0508G	50-54	170	1	.6	15
0510	50-54	170	10	-	-
0510G	50-54	170	10	.6	15
1410	144-148	160	10	-	-
1410G	144-148	160	10	.6	15
1412	144-148	160	30	-	-
1412G	144-148	160	30	.6	15
2210	220-225	130	10	-	-
2210G	220-225	130	10	.7	12
2212	220-225	130	30	-	-
2212G	220-225	130	30	.7	12
4410	420-450	100	10	-	-
4410G	420-450	100	10	1.1	12
4412	420-450	100	30	-	-
4412G	420-450	100	30	1.1	12

- Linear (all mode) RF power amp with automatic T/R switching (adjustable delay). Amplifier useable with drive powers as low as ½ watt.
- Receive preamp option, featuring GaAs FETS (lowest noise figure, better IMD). Device NF typically .5 dB.
- Thermal shutdown protection incorporated
- Remote control capability built-in
- Rugged components and construction provide for superior product quality and performance
- All models include a complete operating/service manual and carry a factory warranty on all components
- Designed to ICAS ratings, meets FCC part 97 regulations
- Approximate size is 2.8 x 5.8 x 10.5" and weight is 5 lbs.

Specifications/price subject to change

1. Models with G suffix have GaAs FET preamps. Non-G suffix units have no preamp.
2. Covers full amateur band. Specify 10 MHz Bandwidth for 420-450 MHz Amplifier.

★*SEND FOR FURTHER INFORMATION*★

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DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

DX ANTENNAS FOR 160-10 METERS. Small size, broadband, high performance. Also, antenna parts, Beverage insulators, wire and cable. Low prices and fast service. SASE for catalog. W1FB, Oak Hills Research, POB 250, Luther, MI 49656.

WANTED: Operating or service manual for Gonset 2 meter linear No. E1022. Uses 2826 tubes. R.J. Heinen, K5TFZ, 1321 Bank Street, Lake Charles, LA 70601. (318) 439 3406.

CABLE TV CONVERTERS & EQUIPMENT: Plans and parts. Build or buy. SASE for information. C & D Electronics, PO Box 1402, Dept. HR, Hope, AR 71801.

YAESU OWNERS: Hundreds of modifications and improvements for your rig. Select the best from fourteen years of genuine top-rated Fox-Tango Newsletters by using our new 32 page Cumulative Index. Only $6 postpaid (cash or check) with $4 Rebate Certificate creditable toward Newsletter purchases. Includes famous Fox-Tango Filter and Accessories Lists. Milt Lowens, N4ML (Editor). Box 15944, W. Palm Beach, FL 33416. Telephone (305) 683-9587.

R-390A Receiver: $195 checked; $115 reparable. Parts, tubes, sections. Info SASE. CPRC-26 six meter transceiver (see HR, March 1985) $17.50 apiece, $32.50 pair (add $4.50/unit shipping). *Baytronics, Box 591, Sandusky, OH 44870.* 419-627-0460 evenings

COLLINS WANTED: Inoperative 75S3C or 32S3 for parts. Mike Palmer, K5FZ, 16707 Creeksouth, Houston, Texas 77068. (713) 444-7737

Foreign Subscription Agents for Ham Radio Magazine

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Ham Radio England
c/o R.S.G.B.
Alma House
Cranborne Road
Potters Bar
Herts EN6 3JW
England

'SUPERLOG' Tandy 1000 and PC compatibles. GW-BASIC $19.95 diskette. TRS-80 Model 4/4P. BASIC or ML. $29.95. KF6CQ, 7438 Kester Avenue, #8 Van Nuys, CA 91405.

SOME QSLs 100 for $6.25. Sample SASE. WB2EUF, Box 708, East Hampton, NY 11937.

B&K Precision solid state signal generator, model E-200D, 100 kHz to 220 MHz $75.00. 4CX1000, factory sealed $250.00 or trade for IC-02AT. W7TZO, (503) 267-6064.

TUBES: 4D32 Raytheon $35.00 each. 4-400A $45.00, 5933WA $5.00, 12AT7 $1.00, 12AX7 $1.00. Inquire on others. Levy, 101 East Driftwood #44, FBG, TX 78624. Tel 512-997-2534. W5QJT.

MARCO: Medical Amateur Radio Council, Ltd. operates daily and Sunday nets. Medically oriented Amateurs (physicians, dentists, veterinarians, nurses, physiotherapists, lab technicians, etc.) invited to join. Presently over 550 members. For information write MARCO, Box 73's, Acme, PA 15610.

WANTED: DX 200 or DX 302 or R7. Must be in working order and good condition. 1116 Bachman Drive, Lexington, SC 29072.

IBM-PC RTTY/CW. New CompRtty II is the complete RTTY/CW program for IBM-PC's and compatibles. Now with larger buffers, better support for packet units, pictures, much more. Virtually any speed ASCII, BAUDOT, CW. Text entry via built-in screen editor! Adjustable split screen display. Instant mode/speed change. Hard-copy, diskcopy, break-in buffer, select calling, text file transfer, customizable full screen logging, 24 programmable 1000 character messages. Ideal for MARS and traffic handling. Requires 256k PC or AT compatible, serial port, RS-232C TU $65. Send call letters (including MARS) with order. David A. Rice, KC2HO, 25 Village View Bluff, Ballston Lake, NY 12019.

ELECTRONIC ENCLOSURES: Attractive, reasonably priced. Aluminum top and sides. Walnut stained solid wood ends. Variety of sizes and custom sizes available. Sample $5.00 Information free. *Energy Engineering, Custom Division, Rt. 4, Box 330, Fayetteville,* AR 72701.

HAM RADIO SETUP in Paradise, CA. 2 B.R. with attached studio apt. 1.8 acres and large barn. Tower - 6 el. DXX Beam - 1500 ft. elev. $135,000.00. Please call (916)-877-8300.

NATIONAL RADIO EQUIPMENT manual list or NCL-2000 parts kits. SASE. Maximilian Fuchs, 11 Plymouth Lane, Swampscott, MA 01907.

ATTENTION AMATEURS *Send for Free discount catalog.* Amateur Communications, 2317 Vance Jackson, San Antonio, TX 78213. (513) 734-7793.

8877 VHF AMP KITS: HV power supplies, CX600N relays, MuTek LTD front end boards for IC251/IC271, EME newsletter and QRO parts. SASE for new catalog. KB7Q, "Q" Products, 417 Staudaher Street, Bozeman, MT 59715.

ELECTRON TUBES: Radio, TV & Industrial Types. Huge inventory. Send for 80% off tube listing. Call Toll Free (800) 221-5802 or write Box HR, Transleteronic, Inc., 1365 39th Street, Brooklyn, NY 11218 (718) 633-2800.

RTTY JOURNAL—Now in our 34th year. Join the circle of RTTY friends from all over the world. Year's subscription to RTTY JOURNAL, $10.00, foreign $15.00. Send to: RTTY JOURNAL, 9085 La Casita Ave., Fountain Valley, CA 92708.

THE GOOD SAM HAMS invite RV operators to check in the Good Sam Ham net 14.240 *Sundays 1900Z also 3.880 Tuesdays at 2359Z. Net control N5BDN, Clarksville, IN*

IMRA, International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, New York 10538.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

SELL: *Central Electronics Nu-Tune 600-L linear. Yaesu FT-301-D.* FT-301 S. Century-21, Digital. Century-21 Analog. Keyers. Calibrators. Prices, trades negotiable. Please write for details. WANT Drake MN-4. HW-202. MFJ-910 antenna matcher. VHF wattmeter. Curtis keyer or IC. Mirage B-23. Mocom-10 or other 10 meter FM rig. WA6GER, 3241 Eastwood Road, Sacramento, CA 95821.

DISCOUNT CATV CONVERTERS/DECODERS and Video accessories. Send for free information and prices - it could save you BIG MONEY on your next purchase of these and other CATV items. Easy View, (HR) PO Box 221K, *Arlington Heights, Illinois* 60006. (312) 952-8504. Ask for Rudy Valentine.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. Daily Electronics, PO Box 5029, Compton, CA 90224. (213) 774-1255.

CUSTOM MADE EMBROIDERED PATCHES. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept 301, 4202 N. Drake, Chicago, IL 60618.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

CABLE TV CONVERTERS/DESCRAMBLERS. Guaranteed lowest prices in US. Jerrold, Hamlin, Zenith - - Many others. Lowest dealer prices! Orders shipped within 24 hours! Mastercard, VISA, COD accepted. Free Catalog - call (800) 345-8927 only. Pacific Cable Co., Inc., 7325-1/2 Reseda Blvd, #1016, Reseda, CA 91335. (818) 716-5914.

HOMEBREW PROJECTS. List SASE. WB2EUF, Box 708, East Hampton, NY 11937.

CHASSIS and cabinet kits. SASE K3IWK, 5120 Harmony Grove Road, Dover, PA 17315.

COMING EVENTS

Activities — "Places to go . . ."

OKLAHOMA: October 5. The Salt Plains Amateur Radio Club will hold their annual Ham Social, Salt Plains Lake, Northern Alfalfa County. Covered dish dinner. Talk in on 147.30/90. Inquiries to: Gary Gerber, KB0HH, 511 Lincoln, Anthony, KS 67003. (405) 842-5076.

ARIZONA: September 27. CARA — The Cochise Amateur Radio Association's Flea Market at the CARA Training Facility on Moson Road, Sierra Vista. No charge for tailgating. Free overnight RV camping for club members. Talk in on 146.16/76. For more information: CARA, PO Box 1855, Sierra Vista, AZ 85636.

CONNECTICUT: *September 28.* The *Waterbury ARC will sponsor* a Flea Market, Waterbury State Technical College, off I-84, Waterbury. 10 AM to 3 PM. Admission $2.00 at door. Indoor spaces $10/table and tailgating spaces $5. Dealers, sellers setup 9 AM. Contact Gary Firtick, K1EB, 589 Hamilton Avenue, Watertown, CT 06795 by 9/15/86.

MARYLAND: October 5. The Columbia Amateur Radio Association's 10th annual Hamfest, Howard County Fairgrounds, just west of Baltimore, off I-70. 8 AM to 3:30 PM. Admission $3.00. Spouse and children free. Reserved tables $7.00 prior to September 30. $8.00 after. Outdoor tailgating $3.00 additional. Indoor tailgating $6.00 additional. Food available. Talk in on 147.735/135, 146.52/52. For table reservation and information: Mike Vore, W3CCV, 9098 Lambskin Lane, Columbia, MD 21045. 992-4953.

PENNSYLVANIA: October 4. Pack Rats (Mt. Airy VHF ARC) invites all Amateurs to the 10th annual Mid-Atlantic VHF Conference, Warrington Motor Lodge, Rt 611, Warrington and the 15th annual Pack Rat Hamarama, Sunday, October 5, Bucks County Drive-In theater, Rt. 611, Warrington. Flea market admission $5.00 per car. Selling spaces $6.00 each. Gates open 6 AM rain or shine.

GEORGIA: November 1-2. The Alford Memorial Radio Club of Stone Mountain is sponsoring Ham Radio and Computer Expo '86, Gwinnett County Fairgrounds, 20 minutes NE of Atlanta. 9 AM to 5 PM Saturday; 9 AM to 4 PM Sunday. Admission $4 advance, $5 at door. Forums, awards, VEC exams both days, free cookout Saturday night. Activities for the entire family. Superb dealer facilities, giant undercover flea market. Discount hotel rooms. Free parking, RV sites with full hookups. Talk in on 146.16/76, 449.25/4.25. Information: Alford Memorial ARC, PO Box 1282, Stone Mountain, GA 30086 or call N8LM at 404-925-7615.

NORTH DAKOTA: September 19-21. ARRL Dakota Division Convention sponsored by the Red River Radio Amateurs, Fargo Holiday Inn. Speakers: Dr. Tony England, W0ORE and ARRL President Larry Price, W4RA. Indoor flea market tables $5. Seminars, display, VE exams (no walkins). Breakfast and banquet Saturday. Breakfast and flea market Sunday. Talk in on 146.76 repeater. Registration $7. For information SASE to Red River Radio amateurs, Box 3215, Fargo, ND 58108-3215.

CONNECTICUT: *September 14.* The *Candlewood Amateur* radio Association's annual Flea Market, Danbury Elk's Club, 346 Main Street, Danbury. 9 AM to 3 PM. Dealers 8 AM. Admission $2. Tables $8. Tailgating $5. Talk in on 147.72/12. For table reservations send check or MO to CARA c/o Gene Marino, 27 Valley View Rd, Newtown, CT 06470 or call Gene (203) 426-8852.

CALIFORNIA: September 20. Sonoma County Radio Amateurs fourth annual flea market. 8 AM to 2 PM, Sebastopol Community Center, 390 Morris Street, Sebastopol, 5 miles west of Santa Rosa. Admission and parking free. Tables $5 advance or $7 at the door. Vendor setup 7 AM. VEC exams, radio clinic, exhibits, refreshments. Auction about noon. Talk in on 146.13/73. For tickets and information: SCRA, Box 116, Santa Rosa, CA 95402.

NEW YORK: September 27. The Elmira Amateur Radio Association's 11th annual International Hamfest, Chemung County Fairgrounds, Elmira. Gates open 6 AM to 5 PM. Tickets available at gate or in advance from Steve Zolkosky, 118 Est 8th Street, Elmira Heights, NY 14903. Outdoor flea market, dealer displays. Breakfast and lunch served on premises.

FLORIDA: September 6-7. The 21st annual Melbourne Hamfest sponsored by the Platinum Coast ARS, Melbourne Auditorium. Saturday 9-5 and Sunday 9-4. FCC exams, exhibits, swap tables, ARRL forum, QCWA, tech, MARS and Net meetings, 2m transmitter hunt. Tickets $3 advance, $4 door. For swap table reservations (limit 2 adj.). For exam information SASE to PCARS, POB 1004, Melbourne, FL 32901.

NEW JERSEY: The Maple Shade ARC is sponsoring its First *Annual Hamfest, Saturday, September 20, Maple Shade High School, Coles Avenue, Maple Shade. 8 AM to 2 PM. Admission* $5.00 per carload, includes one tailgate space. Refreshments, tech forums. Talk in on 146.52 and 223.02/224.62 W2MX Rpt. For more information: K3HW, Howard Weinstein, 15 Lakeside Drive, Marlton, NJ 08053 (609) 596-3304.

MISSOURI: September Swapfest, Sunday, September 28, Harvester Lions Club Park, Harvester. Sponsored by the St. Peters Amateur Radio Club. $1.00 per person to look, swap, sell. Free coffee and donuts for early birds. For information: Joe Riordan, KG0K, 2760 Hwy. 40-61, O'Fallon, MO 63366.

ILLINOIS: September 20-21. Peoria Area Amateur Radio Club is sponsoring Peoria Superfest '86, Exposition Gardens, W. Northmoore Road, Peoria. Gate opens 6 AM. Commercial Building 9 AM. Admission $3.00 advance; $4.00 at gate. Children under 16 free. Amateur Radio and computer displays, hugh flea market, FCC exams for all classes Saturday and Sunday. Free bus to Northwoods Mall Sunday. Fukll camping facilities on grounds. Saturday night get-together Heritage House Smorgasboard, 8209 N. Mt. Hawley Road, Peoria. Talk in on 146.16/76, W9UVI. For information and reservations: SASE to Superfest '86, PO Box 3461, Peoria, IL 61614.

CONNECTICUT: *September 21.* The *4th annual* Natchaug ARA Giant Flea Market, Elks Home, 198 Pleasant Street, Willimantic.

9 AM. Dealers setup 8 AM. "Bargins Galore". Free parking. Admission $2. Under 16 free. Inside reserved tables $5 each. At the door $7 each. Tailgaters welcome. Outside space $5 and up. ARRL/VEC exams for all license classes. Talk in on 147.30/90 and 52. For information: Ed Sadeski, KA1HR, 49 Circle Drive, Mansfield Center, CT 06250. (203) 456-7029 after 4 PM.

CALIFORNIA: October 4. Scatcon '86. 9 AM to 3 PM, Cortez Park, 2441 Cortez Avenue, West Covina. Tech sessions. Hands-on packet, satellite and more. Swaps, vendors. Donation $2. Talk in on 147.765/.600. For more information: Bob, N6NGN. (818) 917-6470.

NEW YORK: September 21. LIMARC sponsors ARRL Long Island Hamfair, New York Institute of Technology, Rt. 25A/Northern Blvd, Old Westbury. Tailgating space $5.00. Hams $3.00 admission. Spouses, kids free. Open 7:30 AM sellers. 9 AM buyers. Food, refreshments available. Talk in on 146.85. For more information: Hank Wener, WB2ALW (516) 484-4322 evenings.

GEORGIA: September 28. The 13th annual Lanierland ARC Hamfest, Holiday Hall, Holiday Inn, Gainesville. 8:30 AM. Free tables and inside display area for dealers reserving in advance. Large parking lot for Flea Market. Novice-Extra VE exams beginning 9 AM. Talk in on 146.07/.67. For information and reservations: Paul Watkins, W4FDK, 5435 Mallard Point, Gainesville, GA 30501 (404) 536-8280.

MICHIGAN: September 14. L'Anse Creuse ARC will have its 14th annual Swap and Shop, L'Anse Creuse High School, Mt. Clemens. 0800-1500. Admission $1.00 advance, $3.00 at the door. Plenty of food and parking. Trunk sales $4.00/space. Inside tables $8.00 each. For Tickets and table reservations SASE to Maurice Schuttecatte, N8CEO, 15835 Touraine Ct, Mt. Clemens, MI 48044. (313) 286-1843.

ONTARIO: September 20. Packet Radio Symposium co-sponsored by the Hex-9 Group of the Barrie ARC and Georgian College, Barrie. Flea market in the morning. Guest speakers Harold Price, NK6K, a director of AMSAT and Ed Jackson of Buffalo. Admission $5.00. Talk in on 146.25/146.85 VE3LSR. Inquiries: Hex-9 Group, Box 151, Orillia, Ontario, Canada L3V 6J3.

MASSACHUSETTS: The Chelsea Civil Defense will sponsor Amateur Radio evening classes at Chelsea High School starting September 11, 1986 for those wishing to obtain a Novice, Tech/General license. There will be a minimal fee for materials cost only. Share in a great hobby! For more information: Frank Masucci, 136 Grove Street, Chelsea, MA 02150. Please include your phone number.

1986 "BLOSSOMLAND BLAST" Sunday, October 5, 1986. Write "BLAST", PO Box 175, St. Joseph, MI 49085.

CALIFORNIA: FCC exams, Novice Extra. Sunnyvale VEC ARC. (408) 255-9000 24 hour. 73, Gordon, W6NLG, VEC.

MASSACHUSETTS: The MIT UHF Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday, September 17, 7 PM, MIT Room 1-134, 77 Mass Avenue, Cambridge MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-0160/646-1641 or Craige Rodgers (617) 494-1986. Exam fee $4.25. Bring copy of current license (if any), two forms of picture ID and completed form 610 available from FCC in Boston (223-6609).

MASSACHUSETTS: September 28. The Wellesley Amateur Radio Society's annual outdoor Flea Market, Wellesley Senior High School parking lot, Rice Street, Wellesley. 9 AM to 2 PM. Admission $1.00 for buyers and $2.00 for sellers. Light refreshments available. Talk in on 147.63/03 repeater.

NEW YORK: October 5. Electronics Fair and Giant Flea Market, Yonkers Municipal Parking Garage, Nepperhan Avenue and New Main Street. 9 AM to 4 PM. Satellite TV, SSTV, Amateur Radio, Computers and more. Giant Auction 2 PM. Unlimited free coffee all day. Admission $3.00. Children under 12 free. Sellers $7.00 per parking space, one admission. Bring tables. For information call (914) 969-1053.

NEW YORK: September 6. Ham O Rama & Computerfest '86. Niagara Falls International Convention Center, Niagara Falls. 7 AM to 5 PM. Displays, tech programs, computer displays, FCC exams. All enclosed flea market. Sell from your car! General interest programming including tours of the Falls. Registration $3.50 advance by 8/20/86. $5.00 at the gate. Inside tailgating $5.00 for 8' x 20' area. For more information: Nelson Oldfield, W2ZSJ, 126 Greenaway Blvd, Cheektowaga, NY 14225.

NEW MEXICO: September 27 and 28. Camp Stoney, 8 miles east of Santa Fe. Saturday ARRL/VEC exams. Free camping at Camp Stoney Saturday night. No hookups. Sunday, tailgate flea market. Registration $5.00 for adults and $2.00 for kids under 12. Includes Sunday BBQ. Talk in on 146.22 82 repeater of 146.52. For more information SASE to Alan Hill, N5BGC, 2020 Calle Perdiz, Santa Fe, NM 87505.

MICHIGAN: October 12. Ham Fair '86, Michigan National Guard Armory, 2500 S. Washington Avenue, Lansing. 8 AM to 3 PM. Dealer sales, swap shop, handcrafted items, FCC exams starting 1 PM. Register for exams by September 12. Donation $3.00 adults. Handicap facilities on premises. For information and reservations: Rowena Elrod, KA8OBS, 111 Lancelot Place, Lansing, MI 48906. (517) 482-9650.

PENNSYLVANIA: September 20-21. York Hamfest sponsored by the York ARC, Keystone VHF Club, Penn Mar RC and Hillstop Transmitting Association. York Fairgrounds, Rt 74, NR corner of the city. Seminars, tailgating, displays, banquet and FCC exams Saturday. Tailgating and displays Sunday. Registration $3.00 per day or $5.00 both days. Women and children under 12 free. Banquet $10 per person advance registration only. Tailgating $4 per day or $6 for both days. Indoor area tables $5 and up per day. York Hamfest, Box W, Dover, PA 17315.

NEW JERSEY: October 4. The Orange County ARC will hold its Hamfest and Auction, John S. Burke Catholic High School. 9 AM to 3 PM. Tailgating $3.00. Setup 8 AM. License exams starting 9 AM. Admission $3.00. Tables $7.00. Talk in on 146.76 Rpt. and 146.52 simplex. For more information call Bob, WB2ENA (201) 767-6698.

MICHIGAN: September 20. GRARA Swap and Shop, Hudsonville Fairgrounds, Hudsonville. Open 8 AM. Admission $3.00. Reserved tables $4.00. For information and reservations: Larry Kozal, K8PUJ, 864 Coldbrook NE, Grand Rapids, MI 49503. (616) 459-8722.

ILLINOIS: September 21. The Chicago ARC will hold an Open House in conjunction with its 60th anniversary. 10 AM to 8 PM at the North Park Village, 5801 N. Pulaski Avenue, Chicago. Special event station on 20 and 40m SSB will operate during those hours. All hams and those interested in Amateur Radio are invited. For information call (312) 545-3622 or write CARC, 5631 W. Irving Park Road, Chicago, IL 60634.

OPERATING EVENTS

"Things to do . . ."

September 4-7: The Stu Rockafellow Amateur Radio Society of Plymouth, MI will be celebrating their 25th anniversary, in conjunction with the Plymouth Fall Festival, operating radio station W3NJH. A certificate will be issued for QSL and SASE. QSL via W8NJH or WD8IAE CBA.

September 6: AT&T Bell Labs Whippany ARC will operate W2TW, 1300Z to 2200Z commemorating its 30th anniversary. Operation in lower portions of 10 80 General phone bands. For a QSL send SASE and QSL to WB2OOO, Rick Anderson, 243 Mountain Avenue, Murray Hill, NJ 07974.

September 6: The 160 Meter Bulletin SSB Contest. 001Z to 2400Z September 7. Single and multi-operator classes. Send logs to 160 Meter Bulletin c/o R. Koziomkowski, KA1SR, 5 Watson Drive, Portsmouth, RI 02871.

September 6-7: The Radio Association of Erie (W3GV) will commemorate Commodore Perry's victory at the Battle of Lake Erie during the War of 1812. 9 AM to 9PM Saturday and 9 AM to 5 PM Sunday local time. A special attempt will be made to contact like vessels of that period and other Great Lakes Historical Vessels Sunday at 1 PM. Special QSL and historical information on Flagship Niagara via W3GV, POBox 844, Erie, PA 16512 or W3QSL Bureau for DX Stations. Please send business SASE.

September 13, 14 and 20, 21: The Valley of the Moon ARC will operate special event station N6KM to commemorate writer, Jack London, author of "Call of the Wild" and "The Sea Wolf" from his home, Wolf House in Jack London State Park, Glen Ellen, California. A beautiful 8x11 certificate suitable for framing, with London's picture and history, is available for a QSL card and $1.00 sent to VOMARC, 358 Patten Street, Sonoma, CA 95476. For an unfolded certificate, please send 9x12 SASE.

September 27: California QSO Party. Sponsored by the Northern California Contest Club. Stations outside California work as many CA stations in as many CA counties as possible. Stations in CA work anyone. CA stations send QSO number and county. Outside stations send QSO number and state/province/country. All logs and summary sheets must be send to NCCC c/o Gary Caldwell, WA6VEF, 1830 Polk Street, Concord, CA 94521 by November 1, 1986.

September 7-14: Southern Counties ARA will operate special event station K2BR during the Miss America Pageant, Atlantic City, New Jersey. QSL SASE via SCARA, Box 121, Linwood, NJ 08221.

September 6: The West Alabama Amateur Radio Society will sponsor a Special Event Station in honor of college football and its greatest coach, Paul "Bear" Bryant. Listen for WD4DAT from 1300Z to 2300Z. To receive the handsome 8x11 certificate SASE with your QSL card to WAARS Special Event, PO Box 1741, Tuscaloosa, AL 35403.

September 28: The Mahoning Valley ARA will operate W8QLY from Boardman Park during the Annual Rotary Octoberfest Celebration. Freq: 40 and 20 phone; 145.01 packet. For a special QSL certificate, send standard SASE to MVARA Octoberfest Station, PO Box 2950, Youngstown, Ohio 44511.

W.E.C.A. (Westchester Emergency Communications Association) initiates Equipment/"Elmer" Banks to help new hams get on the air. More information write the club at PO Box 131, North Tarrytown, NY 10591.

DX CERTIFICATE. Nigerian Amateur Radio Society announces its "NARS at 25 Award." To encourage more contact with Nigerian Amateur Radio stations. DX stations need to work only 5 Nigerian ham stations to obtain this DX certificate. Send proof of contact plus $5.00 to NARS at 25 Award, PO Box 2873, Lagos, Nigeria, West Africa.

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SAY YOU SAW IT IN HAM RADIO

THE GUERRI REPORT

by Ernie Guerri, W6MGI

circuit boards keep pace with electronic developments

Many of us may view printed circuits (PC's) as one of the necessary, but less glamourous aspects of modern circuit design. But the people whose job it is to keep pace with developments in circuit evolution have done some rather marvelous work in the PC area. For example, take a look at one of those plug-in cards for your personal computer — you'll see 50 chips, a ton of components, thousands of interconnects, and hundreds of circuit lines on *six* layers of PC! Chances are you're getting all this for a couple hundred bucks.

Manufacturing modern microwave PC's requires a degree of magic just slightly removed from potions using bat wings and butterfly ears.

Because of the very small dimensions associated with frequencies above 2 GHz or so, the stability and dielectric characteristics of the substrate become a limiting consideration. Teflon and ceramic dielectrics are used extensively — but these are very expensive materials and may cost 50-100 times as much (per square unit) as conventional commercial-grade board materials. Because of the dielectric and radiation losses associated with microwave frequencies, it's not uncommon to have the dielectric bonded to a metallic structural member. This provides shielding, thermal conductivity, and all-important dimensional stability. The copper foil used in such applications is electrolytically formed and may be only 50 millionths of an inch thick (remember the "skin-effect" at very high frequencies). Transmission lines, filters, inductors, etc. may have line widths as small as one thousandth of an inch. Since the characteristics of filters at high microwave frequencies will depend on accurate physical dimensions, great care goes into the selection of the *piece* of PC material that will be used for a given design. At these frequencies it's not unusual for a designer to finalize his design only after he has procured enough material to assure that he can make all of the circuits of a given type that he expects to produce.

Next time you pick up a modern complex circuit card, remember that the card itself is one of the miracles of our current revolution.

signal encryption techniques reach maturity

Recent legislative activities aimed at preventing the reception of certain commercial signals may be less relevant in light of the technical means for security employed by many electronic communications systems. In the early 1970s a data encryption protocol called DES-Date Encryption Standard was developed. This technique is used to protect computer records, financial transactions, diplomatic traffic, and so on. More recently, the operators of broadcast satellites have begun using DES subsets to encode entertainment signals for C-Band satellites.

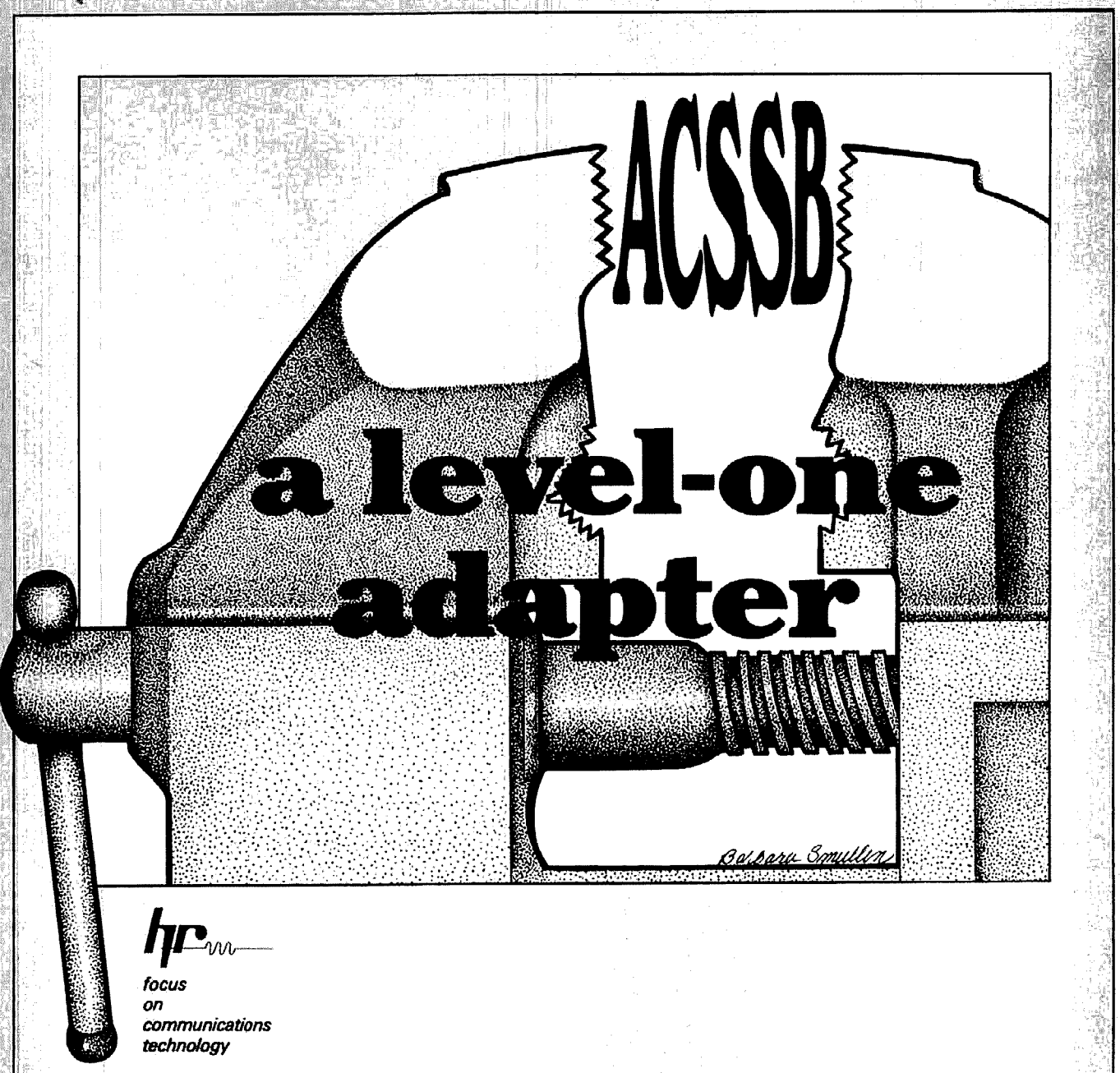
The basic technique is along the following lines... The data is organized into 64 bit words whose right and left 32 bits are swapped in accordance with a coded scheme. The code is a 56 bit word which organizes the structure and rate at which the 32 bit words are swapped. Additional logic operations are performed to compare the right and left hand words. If they don't match in a certain way, one of the words is replaced and a new sequence generated. This process can be extended to sevral levels, and the code can be changed almost continuously. If you think for a moment about the size of number like 2^{32} times 2^{56} you will quickly see why this system has substantial security — the number of possible combinations is astronomical. At this time there are no validated reports of any person or organization having successfully developed a techinque for decoding DES data without a prior knowledge of the code.

Several companies are already producing both encryption and decoding systems using the DES standard. Although the cost is still a bit high for the very low end of the business communication market, prices are coming down. As it has done for us in so many other areas, the TV market is providing the impetus for mass production of DES decoders. One can only hope that those commercial users who would support highly restrictive legislation on the use of the airwaves, can be equally enthusiastic about using modern technology to protect their customer interests.

The possible amateur uses off these advanced coding techniques is not clear, since encryption is prohibited in amateur communications. However, the basic *techniques* (*and especially* the custom chips) might eventually make low cost audio and video digitizers available to the amateur service. This may made possible considerable improvements in spectrum utilization through data compression and frequency/time multiplexing.

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publisher

Rich Rosen, K2RR
editor-in-chief
and associate publisher

Dorothy Rosa, KA1LBO
assistant editor

Joseph J. Schroeder, W9JUV
Alfred Wilson, W6NIF
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editorial production

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director of advertising sales

Dorothy Sargent, KA1ZK
advertising production manager

Susan Shorrock
circulation manager

Therese Bourgault
circulation

cover art:
Barbara Smullen

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REFLECTIONS

a monument to the past?

At the latest ARRL board meeting, the Board of Directors voted to approve in principle the construction of an Amateur Radio museum and visitors' center in Newington. As proposed, the new facility would include 6500 square feet of exhibit space, with several Ham stations representing the progress of Amateur radio through the years on display. There would be a member's library, a lounge, a film viewing room, a small gift shop, and several other sundry rooms — plus extensive renovations for W1AW. The facility's opening would coincide with festivities marking the League's 75th Diamond Anniversary in 1989. Total cost: an estimated $2.7 million.

What comes immediately to mind upon hearing this proposal is that the Board has established a set of priorities that appear to be contrary to the needs of the hobby. By this I mean that we have a far greater need to ensure our future than we do to preserve the past.

This is not to say that the League is not thinking about the future of Amateur Radio. At the October, 1984 Board meeting, two optimistic goals were set: one, to recruit 50,000 new hams per year for the next five years and two, to increase ARRL membership by 25,000 by the end of 1985 and by 20 percent per year thereafter (see *QST*, 12/84, p. 52). At the most recent board meeting, however, the League's executive vice president reported that League membership has grown from 129,698 to 139,910 — an increase of approximately 8 per cent. *The same report disclosed only a 2.3 percent increase in the number of hams during 1985. (*QST*, 9/86, p. 63). So while the League has done a fair job increasing its membership, it hasn't come close to meeting its number-one goal: increasing the Amateur ranks.

On the positive side, several good things are happening. Most notably, the Archie comic book — a $20,000 project, half-funded by the ARRL, promoting Amateur Radio to youngsters — is nearing completion and should be ready for distribution as you read this. A League-sponsored proposal for the expansion of Novice privileges awaits consideration by the FCC. The VEC program seems to be working successfully.

Bringing the museum and visitors' center to completion will consume a tremendous amount of time in planning, fund-raising, construction, and installation and maintenance of the exhibits. Once established, operating costs will increase the League's budget as new employees are added to meet the increased needs of the larger physical plant. Over the next few years, the executive vice president and ARRL staff will spend an inordinate amount of time on this project — time that would be better spent working to meet the goals set by the Board in 1984.

Can you imagine what could be done with a $2.7 million "war chest?" That's more than 270 times the ARRL contribution to the Archie project alone! The scope of projects that could be instituted with this kind of funding is mind-boggling.

There'd be money to support a speakers bureau to send dynamic, persuasive representatives of Amateur Radio into schools, scout meetings, and other places where potential hams congregate. Attractive educational materials based on an upbeat theme (how about *Amateur Radio — 75 years young and looking for YOU!?*) could be professionally produced and distributed. Movies, public service announcements, posters, and other promotional materials could be made available. How about producing a set of low-cost, instructional videotapes (or disks for popular home computers) for self-teaching or small-group study of the fundamentals of Amateur Radio — a la Tune In The World? I've just scratched the surface; there are innumerable ways this money could be used to advance the cause of Amateur Radio.

A far better achievement to mark the 75th Diamond Anniversary of the ARRL would be to meet or exceed the goals established by the Board in 1984. Preserving the past is an admirable goal indeed — but not at the cost of the future!

Craig Clark, N1ACH
Assistant Publisher
ARRL Life Member

*Of this number, only 126,281 are full members; the larger figure includes associate members and others who subscribe to QST but do not belong to the ARRL (*ARRL Letter*, 7/7/86, p. 2).

presstop de W9JUV

AUTOMATIC TRANSMITTER IDENTIFICATION HAS BEEN PROPOSED BY THE FCC in a Notice of Proposed Rule Making released August 8. Though General Docket 86-337 is directed at video satellite uplink stations in response to "Captain Midnight" having broken into an HBO movie last spring, of more importance to Amateurs and other services is a Notice of Inquiry also included, asking whether an Automatic Transmitter Identification System (ATIS) shouldn't be required for all transmitters used by the various FCC licensed services!

Many ATIS Are Already In Use In Police and other U.S. communications systems. In the mid 70s, at the height of the CB craze, ATIS was also proposed for that service as a means of curbing the rule-breaking so rampant at that time. It was not adopted.

No Specific Suggestions For General ATIS Use were made by the FCC in its NOI. Instead, they ask only general questions (about the need for ATIS, its applicability, implementation, timing, format, and similar subjects) and seek user response to the proposal. They also solicit comments on adding ATIS to existing radios, though the resulting logistical and financial burden would make that possibility rather unlikely. In conclusion, the NOI also suggests that various user groups and other interested parties may wish to form an "Advisory Committee" -- possible with FCC support -- to review the many issues involved and devise a mutually acceptable ATIS format.

Comments On The ATIS Proposal, General Docket 86-337, are due at the Commission on October 20; Reply Comments must be submitted by November 19.

JAS-1, RENAMED "FUJI" BY THE JAPANESE, IS UP. Launched August 12 from Japan's Tanagashima Space Center, the new OSCAR's Mode JA transponder (145.9-146.0 MHz up, 435.9-435.8 MHz down -- inverted) has been working well, though users report deep spin-induced fading because the satellite hasn't yet stabilized. Considerable user dexterity is also required to compensate for large doppler shift (up to 15 kHz) as a result of the satellite's rapid orbital travel. Users are reminded to correct for doppler shift with the transmitter's VFO rather than the receiver.

OSCAR 10 Continues To Come On And Shut Down as programmers try to determine what parts of the bird's damaged memory still function, then write software that will provide reasonably reliable operation. At presstime OSCAR 10's future is unknown.

FEDERAL RULES LIMITING RF EXPOSURE SEEM ALMOST CERTAIN, with three government agencies now involved in their development. The Environmental Protection Agency is leading the effort, with the Food and Drug Administration participating and the FCC also involved to some degree.

Exposure Limits Suggested By ANSI (American National Standards Institute) have long been voluntary in the U.S. The RF exposure limits in many other nations are much stricter, however, and it appears that the EPA may adopt a standard that is five times stricter than the ANSI limits. U.S. commercial radio interests, though concerned about the proposed EPA rules, appear likely to go along with the federal effort in order to forestall further development of inconsistent individual state RF exposure regulations.

Though Amateur Radio Had Been Exempted from much previously proposed RF exposure rule-making, it appears likely this time all services will be covered. What effect all this will have in the real world of transmitters and antennas remains to be seen.

RESPONSIBILITY FOR ALL AMATEUR EXAM QUESTIONS WAS DELEGATED to the Amateur Radio community through the Volunteer Examiner Coordinators in an FCC Report and Order released at the Commission's Washington VEC Conference August 9. The Commission's decision on PR Docket 85-196 had been opposed by the ARRL out of concern for exam uniformity; the FCC said "VECs are capable of satisfactorily performing the task and are in the best position to do it in their role as coordinators." However, the Commission did agree with the League that exam standardization was important, and an important result of the VEC conference was the establishment of a "Committee for Question Pool Maintenance" made up of N4BAQ (Western Carolina VEC), WB9IHH (ARRL VEC), and W6NLG (Sunnyvale VEC). Questions for Amateur exams can come from many sources, but their acceptance for the exam pool will be by this committee.

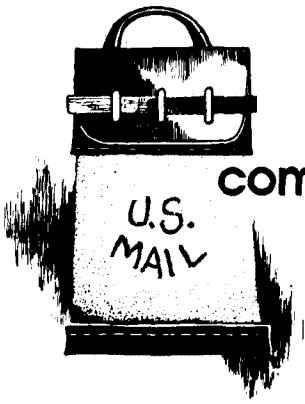
A New Class Of Volunteer Examiner, The "Preparing Volunteer Examiner," was another development of the Report and Order. Present VEs will now be known as "Administering VEs" (AVEs), while the PVEs will be individuals who submit questions for or prepare exams for use by the VECs or AVEs. PVEs will also be permitted to market packaged exams to AVEs.

"CARE KIT," A Complete Novice CW and Written Exam Package, is now available free to VEs and clubs on a first-come, first-served basis from CARE, Box 688H, Glenview, IL 60025.

STILL ANOTHER VERSION OF THE COMMUNICATIONS PRIVACY ACT was introduced in the Senate just before summer recess, but didn't make it to the Senate floor before adjournment. This new version permits monitoring Land Mobile transmissions and reduces some penalties for violations of the Act. Added, however, is a new 1 to 10-year prison term and up to $250,000 fine for willful interference to satellite communications ("Captain Midnight" vs. HBO).

Whether The Summer Recess Will Result In Still Further Changes in the Act is uncertain, though the pressure to further temper its repressive aspects continues.

A NEW "INTERFERENCE HANDBOOK" HAS BEEN ANNOUNCED by the FCC. The extremely well done 64-page manual covers essentially all kinds of TVI and RFI and also includes information on obtaining assistance from most home entertainment electronics manufacturers. It's available free for the asking from any FCC Field Office.



comments

YLs in Amateur Radio

Dear HR

Though once considered a man's hobby, Amateur Radio is now seeing a growing influx of YLs. Although the theory and code requirements are the same for both genders, some are voicing their concerns about the growing number of licensed YLs and the long-term effect on "their" repeaters. Many have expressed fears that the bands will soon be overrun by inexperienced YLs who will turn 2 meters into a glorified CB band filled with meaningless gossip.

To be adequately prepared for entering the world of repeater operations, I — a newly licensed technician — first turned to the FCC manual on Rules and Regulations. Although a basic description of repeater operation was given, it was suggested that the local repeater be monitored to observe the accepted operating procedures. So, to avoid any further accusations of engaging in meaningless gossip, I monitored the local repeater for approximately four hours and compiled a list of accepted topics currently being discussed by the OMs and noted some basic acceptable operating procedures:

- You may talk about baseball, football, cheerleaders and the girls in the stadium.
- You may talk about your favorite restaurant, the foods you are most hungry for, and how long it has been since your last meal. You can even exchange recipes.
- You may talk about car repairs, home repairs, or even vacations. But if you're a YL, don't mention household chores such as laundry or you'll be accused of trashing the air waves with CB type communications.
- You may talk about your wife's moods, her dermatological problems, and even her dental problems.
- You may discuss gardens, grandchildren, and pets or even "the old lady" back at the home QTH. (I thought Q signals were for non-verbal communications?)
- Hobby talk is acceptable if the hobby happens to be golf, cycles, fishing, guns, or photography.
- If it's Sunday, you may discuss church services, choirs, attendance, and even have "roast minister."
- You may comment on road construction or complain about the car in front of you or the car behind you. (But if you're a YL, don't mention that you're mobile chaffeuring children or you'll be accused of misusing the repeater.)
- OMs can ask: "What's up?" or "What's happening?" or "How are things with you?" But if you're a YL, you'd best refrain — because you'll be accused of attempting meaningless gossip and it will be implied that you should stick with the telephone.
- You can certainly talk about the weather . . . .
- If something is funny, don't laugh or you'll be accused of being tacky. A verbal "HI HI" is acceptable, however. Don't giggle — you'll be accused of flirting.
- I've also observed that it's apparently acceptable etiquette for an OM to sneeze, cough, clear his throat, or even burp into a keyed mike.

I don't mean to cause undue strife in the Amateur Radio community or cause a battle between the sexes, but honestly, give us YLs a chance! We worked just as hard as you did to get our licenses and we hold the same pride in our accomplishments (and the same license) that you do. We too are unimpressed with the CB mentality and want the Amateur Radio bands to maintain the high standard of operation that first attracted us to this service.

So, OMs, please listen first before you cast stones.

(Name Withheld by Request)

increasing Amateur ranks

Dear HR

Recently three local Amateur clubs — the Enfield Radio Amateur Group (ERAG), the Mt. Tom Amateur Repeater Association, and the Pioneer Valley Repeater Association — joined in a venture to increase public awareness and to encourage growth of the Amateur fraternity.

A large local shopping mall allowed us to put on an Amateur Radio display in the central concourse; the mall agreed to provide professionally lettered display signs, assist in the erection of the tower, antennas, and power drops, and also agreed to supply tables and drop-cloths. The radio clubs were required to have insurance coverage.

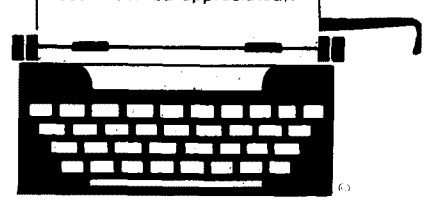
Several displays were set up to demonstrate packet radio, HF communications, traffic handling, and VHF repeater operations. A VCR presentation covered Amateur emergency communications. While several club members manned these exhibits, others talked to the public, answering questions and obtaining the names and addresses of those interested in becoming Radio Amateurs.

During the first day, several hundred people viewed the exhibit and 45 enrolled in Amateur licensing courses offered by ERAG.

Peter J. Bertini, K1ZJH
Somers, Connecticut

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ACSSB:
a level-one adapter

Expansion and compression techniques can yield significant S/N improvement

Everyone who uses communications frequencies is aware of the problems of channel sharing and interference. In many parts of the United States, repeater coordinating councils have declared the 2-meter subbands "full." The CB bands are hopelessly cluttered, and commercial users of VHF/UHF frequencies are searching for a solution to channel overcrowding in the land mobile bands. Obviously, better spectrum management — re-use of existing channels, for example — or new spectrum-saving techniques will be necessary if these problems are to be solved.

One spectrum-saving technique just beginning to be used in commercial communications is ACSSB (Amplitude Companded, or "compandored," SSB). This system consists of a compressor at one point in a communication path (to reduce signal amplitude) followed by an expander at another point (to improve the ratio of the signal to the interference entering the path between compressor and expander). Unfortunately, current implementations of the technology are quite expensive by Amateur standards (over $1200).

ACSSB systems provide the following:

- 4:1 compression of speech during transmission;
- 1:4 expansion during reception;
- Transmitter spectrum equalization (pre-emphasis);
- Receiver spectrum equalization (de-emphasis); and
- Pilot tone reference for AGC control (up to 20-50 Hz fading, typical), automatic frequency control (phase locking), and squelch operation.

The process of compression on transmit and expansion on receive is called "companding" or "compan-

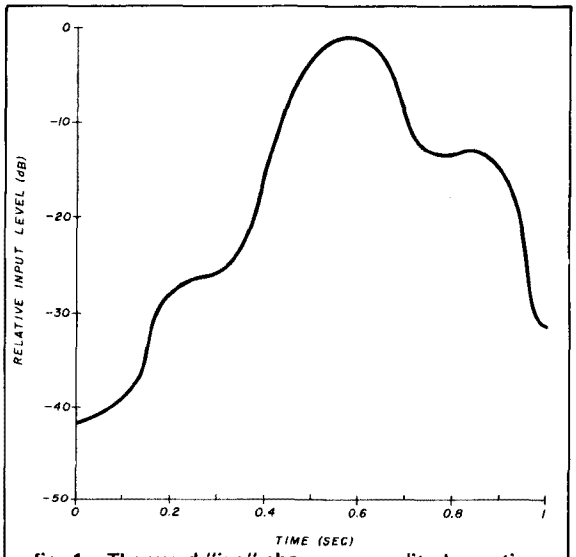


fig. 1a. The word "jug" shown as amplitude vs. time.

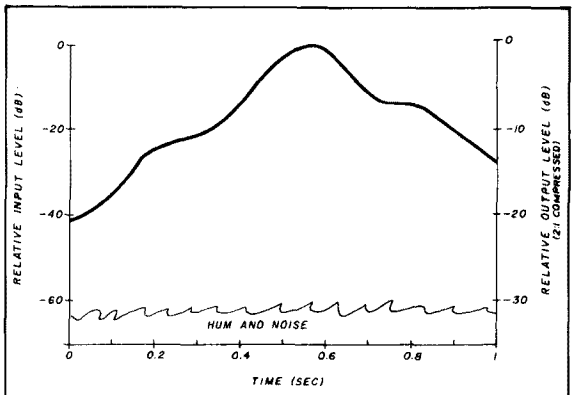


fig. 1b. The word "jug" shown at normal vs. 2:1 compressed dynamics.

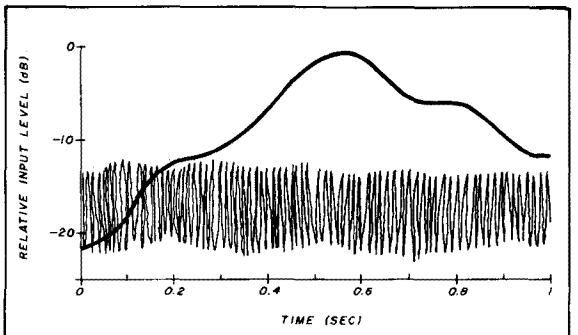


fig. 1c. 2:1 compressed "jug" received 12 dB above the noise.

By James Eagleson, WB6JNN, 15 Valdez Lane, Watsonville, California 95076

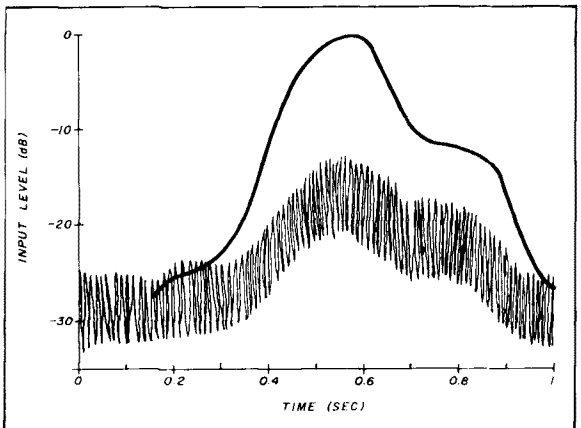


fig. 1d. The signal received in fig. 1c after it is de-emphasized and passed through a 1:2 Expander (2 dB of output change for each 1 dB of input change).

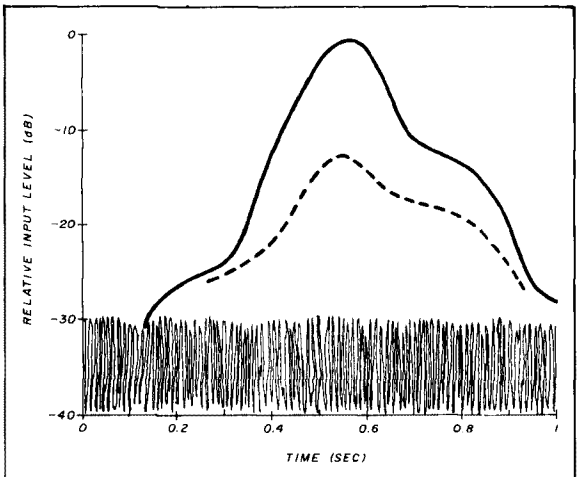


fig. 1e. Direct comparison of FM quieting performance relative to 2:1 ACSSB when both signals arrive at 12 dB SNR occurs much earlier with 2.4 kHz ACSSB than with 20 kHz FM, of course.

doring." For signals above a certain threshold level (5 dB for 4:1 and 10 dB for 2:1 systems), the apparent signal-to-noise ratio will be improved by the compander ratio (either two or four times).

Thus, these systems provide all of the normal conveniences of hands-off operation found in FM systems, but require a channel spacing of only 5 kHz instead of the usual commercial 20 to 25 kHz spacings. (Amateurs use 15 kHz in some instances, but for similar channel-to-channel protection ACSSB channels could be spaced 4 kHz apart). Furthermore, the quieting and capture performance is very similar to that of FM, with ACSSB having slightly improved threshold over FM, although FM has somewhat better transparency.* But $1200 is still $1200!

*Transparency — The ability to pass natural, uncolored sound. Related to SNR, harmonic & intermodulation distortion, tonal balance, transient and phase responses of a system.

level-one circuit

A system offering several of the same performance advantages as commercial ACSSB, but at a fraction of the cost, has been developed by Project OSCAR for use with OSCAR 10. Unlike the commercial versions, which require complete integration of RF and processing circuits, this basic — or Level One — circuit will also work with most SSB transceivers and provide meaningful performance improvements when used on HF through UHF bands.

The Level One ACSSB adapter allows:

- 2:1 compression on transmit (**fig. 1**);
- 1:2 expansion on receive (**fig. 1**);
- Transmitter spectrum equalization (pre-emphasis) (**fig. 2**);
- Receiver spectrum equalization (de-emphasis) (**fig. 2**).

At this time there is no provision for a pilot tone AGC circuit (which could be termed Level Two); however, work is scheduled to provide a Level Two unit that will enhance the Level One adapter by providing a pilot-tone AGC system, squelch, and possibly a second level of companding (yielding a total of 4:1 companding). For simplicity the Level Two circuit will not use phase locking of the pilot, but will be manually tuned instead.

transmit adapter

The transmit adapter (**fig. 3** and **table 1**) has a microphone preamplifier that increases the audio to the proper level for processing. This preamp also provides a moderate level of pre-emphasis — about 4-5 dB at 3000 Hz.

The microphone selected should be low impedance (under 500 ohms). Otherwise the remaining op amp in the 3403 (quad op amp) should be made into a buffer stage in front of the preamplifier stage (**fig. 4**).

As shown in **fig. 2A**, normal distribution of voice energy drops off rapidly above 700 to 800 Hz. The second stage in the transmit adapter equalizes this dropoff by boosting frequencies above 800 Hz to compensate for the rolloff. Thus the audio energy is spread evenly across the entire 300-3000 Hz spectrum (**fig. 2B**). **Figure 2C** shows system response with net or combined equalization.

This provides several advantages. First, it provides more high tones to allow the ALC to function more efficiently (more cycles per second allows faster attack time). Secondly, it reduces the intermodulation prod-

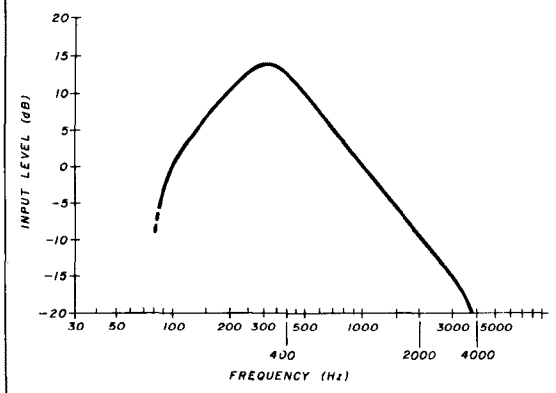


fig. 2a. Typical spectral density of human voice relative to frequency.

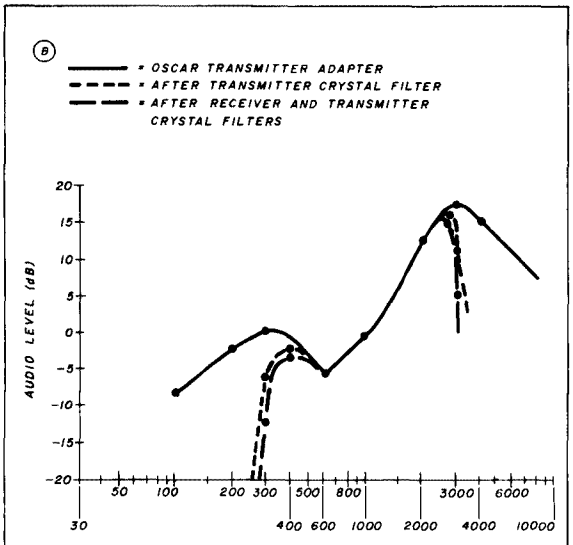


fig. 2b. Basic response of Level One transmission equalizer.

ucts somewhat by lowering the relative level of low voice tones. Thirdly, it raises the important sibilant ("ch," "s," "f") sounds so that they will be above the receiver's noise level entering the radio instead of being 8 to 12 dB below the noise, as would be the normal case at 10 dB signal-to-noise ratio (SNR) without equalization.

The third stage provides 2:1 compression at a syllabic rate. Each syllable is compressed individually so that the typical 40-dB dynamic range of human speech is compressed into only about 20 dB of output level change. This technique, borrowed from telephone long-line services, uses the NE570 or NE571 integrated circuit. Similar techniques are used for Dolby and DBX recording systems.

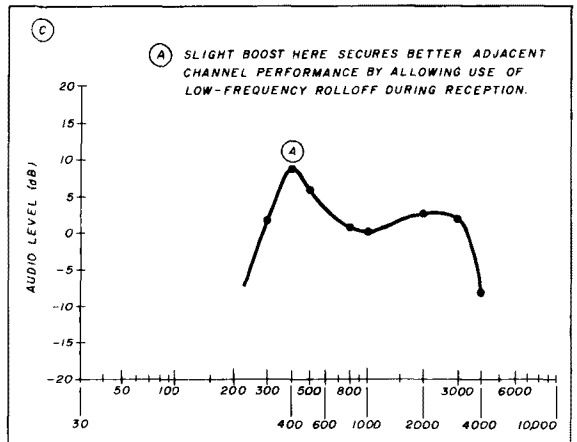


fig. 2c. Equalized audio levels as transmitted (Curve 2a + 2b = 2c).

The advantage of the 2:1 compression used here over normal compression techniques is that the relative dynamic changes in speech levels, though modified, are *maintained after compression*. Most compressors merely suppress loud sounds falling above a certain level. A 2:1 compression ratio generally presents a more pleasing and natural sound to the audio since relative dynamics are maintained.

Furthermore, due to control of each syllable the average power will be raised somewhat. While not as effective as RF clipping, the 2:1 compressor has better harmonic and intermodulation distortion capability.

Finally, the audio is attenuated to microphone audio levels once again then fed to the transmitter. To provide maximum flexibility with the adapter individual switching is provided for each stage. This can be used or not used, as required. A second level of equalization is provided for use with stations that do not have the receive adapter. Normally about 15 dB of boost is used. A switch setting giving only about 10 dB is available also. Of course, 2:1 compression only can also be used, though the slight pre-emphasis, 5 dB, is present all the time.

receive adapter

The receive adapter deprocesses the transmitted signal by restoring the original frequency response and dynamics. A schematic diagram and parts list are provided in **fig. 5** and **table 2**.

A gain stage allows optimizing the levels into the adapter while using normal volume settings on the receiver's volume control. It will also allow adapter use inside the receiver between the detector output and the top of the volume control pot. This latter is the best way to use the adapter since it will eliminate the

table 1. ACSSB Level One transmit adapter parts list.

item	description
	capacitors
C1	1 μF tantalum (TM 1/35)*
C2	1 μF, tantalum (TM 1/35)*
† C3	1000 pF Epoxy Dipped Ceramic (EDC) (272-154)**
† C4	1000 pF EDC (272-154)**
C5	1 μF, tantalum (TM 1/35)*
† C6	1 μF, tantalum (TM 1/35)*
C7	10 μF, tantalum (TM 1/35)*
C8	1 μF, tantalum (TM 1/35)*
† C9	1000 pF EDC (272-154)**
† C10	1000 pF EDC (272-154)**
C11	1 μF, tantalum (TM 1/35)*
C101	1 μF, tantalum (TM 1/35)*
C102	22 μF, tantalum (TM 22/6)*
C103	0.22 μF (TM 0.22/35)*
C104	1 μF, tantalum (TM 1/35)*
C105	0.1 μF(TM 0.1/35)* (272-158)**
C106	10 μF, tantalum (TM 10/25)*
CBY	1 μF, tantalum (TM 1/35)*
	integrated circuits
U1	MC3403 or equivalent
U2	NE570 or 571
	resistors
R1	47 kilohm
R2	10 kilohm potentiometer
R3	1 kilohm
R4	not used
† R5,R6	470 kilohm
† R7	1 Megohm
R8,R9	10 kilohm
R10	36 kilohm
† R11	470 kilohm
† R12	18 kilohm
† R13	180 kilohm
R14-R16	not used
R17,R18	470 kilohm
R103	10 kilohm
R102	47 kilohm
R101	680 kilohm
R104	470 kilohm
R105	1 kilohm (see note)
R106	1 kilohm (see note)
R107	1.5 kilohm
R108	3.3 kilohm
R109	4.7 kilohm
R110	1 kilohm potentiometer
R111	150 kilohm
R112	470 kilohm
	switches
S1,2	DPDT (275-1546)**
S3,4	SPDT (275-625 or 647)**
	voltage regulators
U3	7806 or 78L05
U4	7812 or 78L12
	miscellaneous
box, lid	Pomona 2902
knob	to fit R2 potentiometer

connectors as required

standoffs 4-40 threaded × 1 inch (25.4 mm) long

Note: Select for about 3-6 dB loss through adapter in the out position with $\Delta G = 0$ and drive potentiometer = max.

* JAMECO

** Radio Shack

† 5 percent tolerance, if possible

need for an external amplifier and speaker. It will also provide less distortion by eliminating any receiver audio distortion or hum at the input to the adapter.

The de-emphasis circuit is provided by two stages. The first is a simple 800-Hz low-pass filter, which provides a good compromise between tonal balance and sound quality when the unit is used in the expand mode on noncompressed, nonequalized input signals. The second filter is a 600-Hz band-pass filter, which can also be used for reception of noncompressed signals. Its tonal quality provides more punch to the received audio. When receiving fully compressed and equalized signals, both filters restore the original audio frequency response. Rolloff of low-pitched and high-pitched adjacent channel interference will also result to a certain extent.

Finally, the 1:2 expansion circuit decompresses the incoming audio in the opposite way from that in the 2:1 compressor. That is for each 1 dB of input audio change, the expander provides 2 dB of output change. Since this occurs on each syllable (at about a 100 ms time constant), noise between words will fall to a level that is much lower than the actual SNR of the incoming signal (for example, for 10-dB SNR incoming, the 1:2 expander drops the noise to −20 dB relative to peak audio output).

The expand limiter, fig. 5, biases the expander's control port so that signals below a certain input level will not be expanded. This action reduces the peaky sound caused when the expander is used with noncompressed transmit signals. The value can be selected to suit the user, or a 10-Megohm pot may be used in series with a 1-Megohm resistor to provide continuous control of expansion in place of a switched position.

Since the audio is de-emphasized before applying 1:2 expansion, the SNR enhancement is somewhat better than 2:1. The audio bandwidth after de-emphasis is about 1 kHz out of 2.7 kHz of transmitted audio so that the SNR enhancement due to bandwidth limiting is about 4.3 dB ($10 \log_{10} 2.7/1.0$). Thus an incoming 10-dB SNR signal would be improved to 14.3

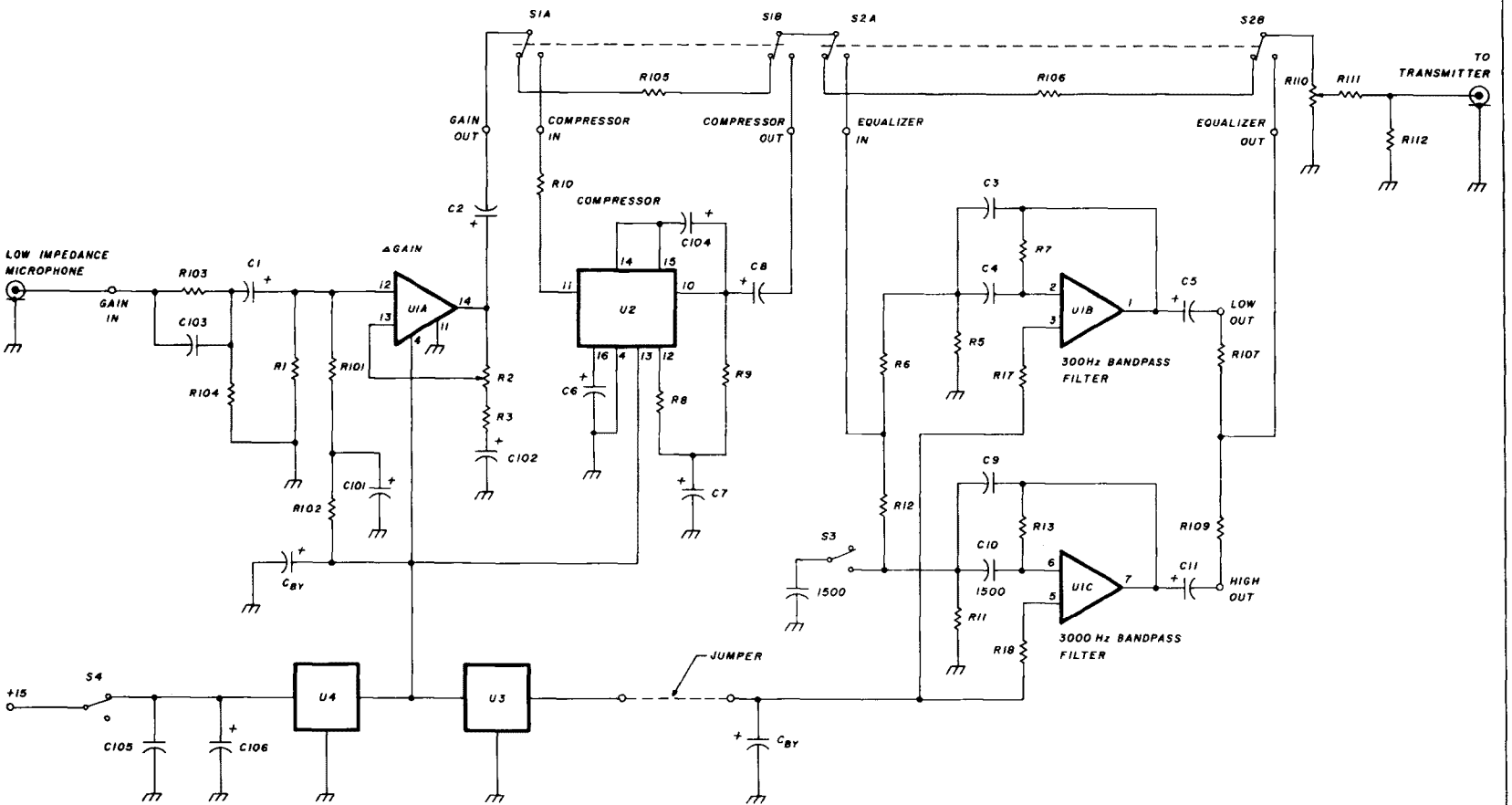


fig. 3. ACSSB Level One transmit adapter schematic diagram.

table 2. ACSSB Level One receive adapter parts list.

item	description
	capacitors
C1	1 μF (TM 1/35)*
C2	22 μF (TM 22/6)*
C3	1 μF (TM 1/35)*
† C4	4700 pF 10 percent (MY 0.0047/100)* (272-155)**
† C5	0.1 μF 10 percent (MY 0.1/100)* (272-158)**
C6	1 μF (TM 1/35)*
† C7,C8	0.01 μF 10 percent (MY 0.01/100)*(272-156)**
C9,C10,C11	1 μF (TM 1/35)*
C12	4.7 μF (TM 4.7/35)*
† C13	C13 1 μF 5 percent (TM 1/35 OK)*
CBY	1μF (TM 1/35)*
C101	10 μF (TM 1/35)*
C102	0.1 μF (MY 0.1/100)
	resistors
R1	10 kilohm
R2,R3	47 kilohm
R4	10 kilohm trim potentiometer
† R5	36 kilohm
R6	(not used)
† R7	36 kilohm
† R8	18 kilohm
R9	47 kilohm
† R10	36 kilohm
R11-R13	(not used)
R14,R15	47 kilohm
R16,R17	(not used)
R18	3.3 Megohm
R19	36 kilohm
RL	(load resistor to match source. 8 ohms or 100 kilohms, typical)
R100	1 kilohm (jumper if more gain is needed)
R101	4.7 kilohm
R102	10 kilohm
Rv	potentiometer, 500 ohms (low impedance in) 100 kilohm (high impedance in) (audio taper)
	switches
SW1-4	DPDT (275-1546)**
SW5	SPDT (275-625 or 647)**
	integrated circuits
U1	MC3403 or eqivalent*
U2	NE570 or NE571*
	voltage regulators
U3	7806 (78L05 OK)* **
U4	7812 (78L12 OK)* **

* JAMECO
** Radio Shack
† 5 percent tolerance, if possible

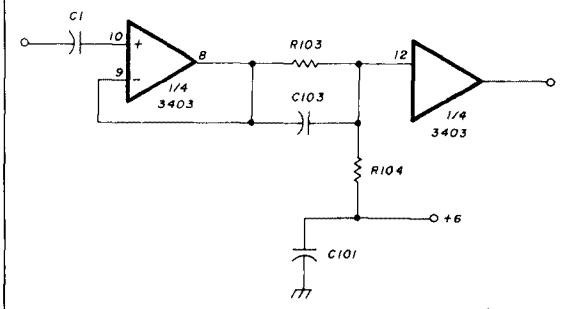


fig. 4. Suggested microphone input circuit to make pre-emphasis impedance insensitive (i.e., for medium/high Z microphones).

dB after de-emphasis, then would be expanded 2:1 to 28.6 dB effective SNR.

In reality, however, some of this SNR improvement, though measurable with instrumentation (such as a VU meter) is lost due to the human ear's frequency sensitivity. The Fletcher-Munson curve* shows a higher sensitivity to mid-range tones than to those above and below roughly 700-1000 Hz. Actual subjective SNR enhancement falls somewhere between 10-14.3 dB improvement over our incoming 10-dB SNR — a good guess being 22-24 dB equivalent SNR after expansion.

CW filter

Though not directly intended for CW operation, it so happens that our lowpass filter, bandpass filter, and expander circuits provide a moderately good CW filter system. The combination lowpass filter/bandpass filter peaks in the 600-1000 Hz region, and the time constants of the expander are about right for normal CW speeds. The expansion provides selectivity about one-half the bandwidth of the bandpass filter/lowpass filter alone but does not cause the ringing often associated with audio filters. On the other hand, any fading present on the signal will be intensified when the expander is in operation.

AGC and fading

Unfortunately, radio circuits are not telephone circuits and noise, fading, and interference are frequently present. While AGC circuits reduce fading significantly, many receivers have AGC overshoot, undershoot, and a control range that is only marginally acceptable. Any such anomalies in receiver AGC performance will be worsened by a 2:1 ratio when used with an expander circuit. Use of the RF gain control

*Also called equal-loudness contours. A group of sensitivity curves showing the characteristics of the human ear for different intensity levels between the threshold of hearing and the threshold of feeling. The reference frequency is 1000 Hz.

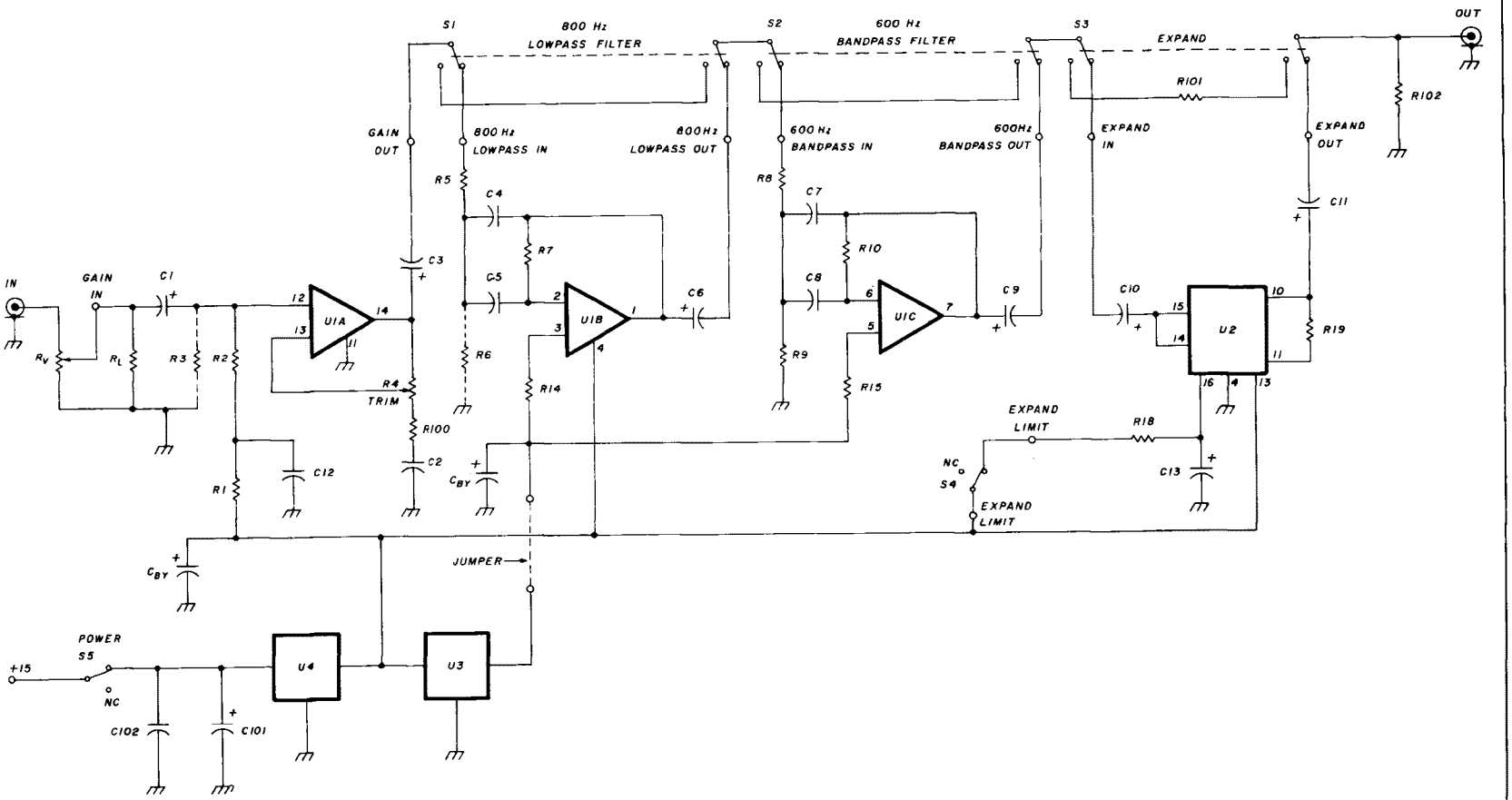


fig. 5. ACSSB Level One receive adapter schematic diagram.

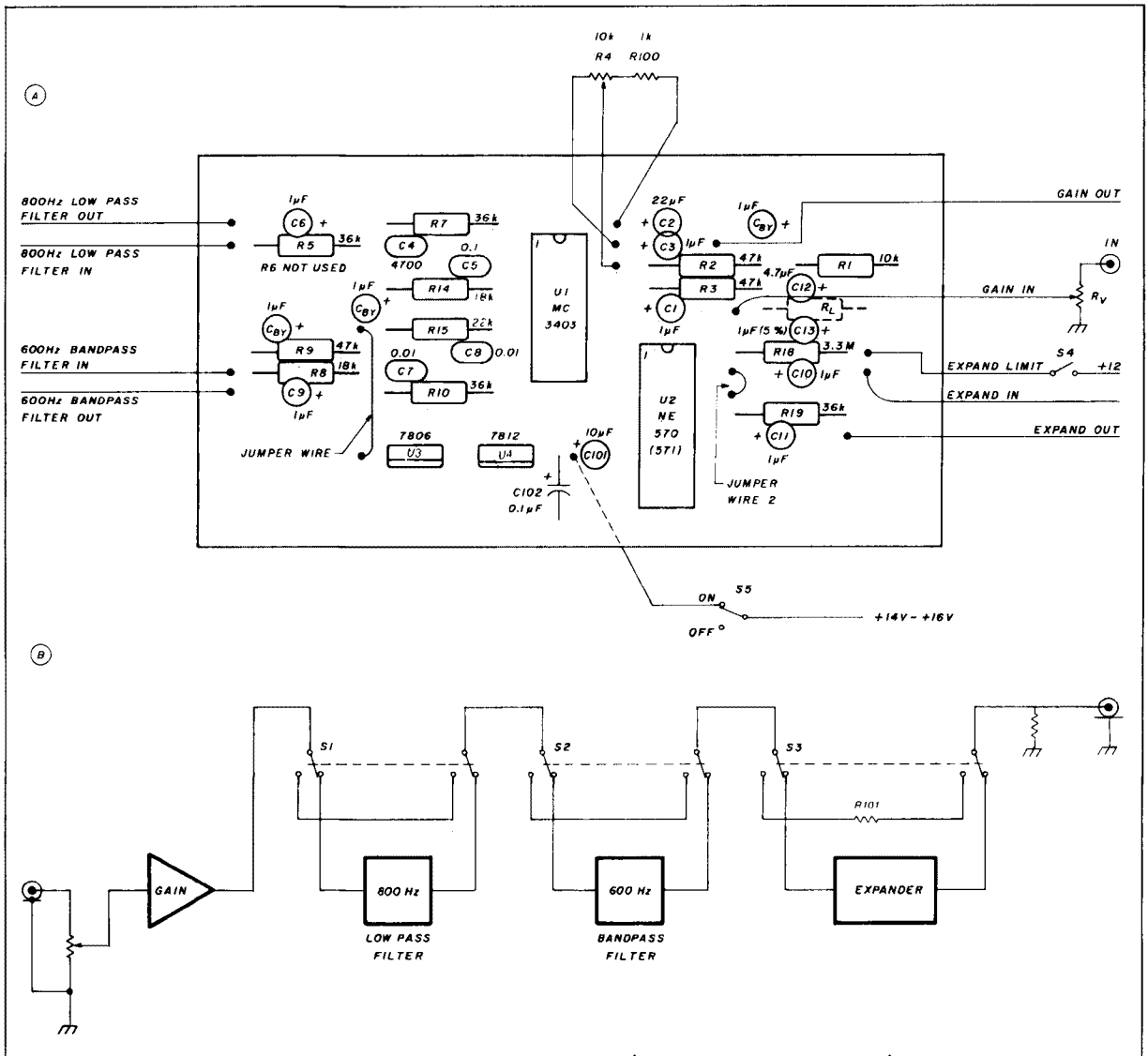


fig. 6. Level-one receive adapter: (A) component placement diagram and (B) block diagram.

and/or fast AGC may be necessary to keep this kind of problem under control.

The Level Two adapter being developed by Project OSCAR should lessen these problems to some extent. It will also provide a tighter control of AGC performance under rapid fading conditions and allow use of carrier-operated relay (COR) and squelch circuits as well as an extra level of 2:1 companding (making a total of 4:1 for the combined system).

Level Two will also allow use of Gateway† stations with automatic frequency locking by adding phase-locked-loop (PLL) circuits to the receiver (Level Three). This would use the pilot tone of Level Two as a known frequency reference. Level Three phase locking can also be implemented by experienced Amateurs with their home equipment, but thorough knowledge of the receiver's internal workings and phase-locked-loop techniques is probably a prerequisite to implementing such circuits.

construction

The OSCAR Level One adapter has no audio output stage. This is partially because there was no room in the Pomona 2901 box for the audio stage and partly because the adapter is really designed to use between

†Gateway stations — sophisticated ground stations that accept several inputs and act as an uplink to the satellite.

using level-one adapters

The effects of compression and expansion are best when used when all parts of the system are completely stable — that is, same gain and frequency response are maintained.

The compressor circuit is the easiest to understand in terms of set-up. It merely needs to be placed on an oscilloscope; microphone gain R2 should be adjusted so that the output at the top of R110 has the same peak value as unprocessed audio. Some adjustment of the values of R105 and R106 may be required if total balance is to be achieved between the various modes of the unit.

This can also be done using the transmitter output by adjusting the drive control, R110, for about half the normal transmitter output on the RF output meter of the transmitter, then adjusting microphone sensitivity, R2, and drive control, R110, until all modes give about the same level on the meter. Again, some adjustment of R105 and R106 might be required.

After this is done, boost the drive control for normal transmitter output. If ALC is used in the transmitter you may be able to boost the output slightly more into the ALC. This may provide slightly more output when using the 2:1 compressor since the peak value of the audio will be maintained more evenly than without the compression (that is, the peak-to-average is better).

You will probably find that the compressor is useful under most conditions to provide more even audio to your transmitter. On the other hand, 2:1 compression alone won't be noticed by most people unless you have high background noise or room echo.

The 10-dB equalization will definitely yield more "brightness" to your signal, thus increasing clarity under weak signal conditions. The 15 dB may be too "bright" without some amount of de-emphasis during reception, however. Then again, under weak signal conditions, either position will help improve clarity.

Note that a slight amount of pre-emphasis is present at all times. This is not extreme, but high-pitched microphones should probably be avoided. Also note that the input circuit assumes a 500-ohm or lower impedance microphone. Values of the input pre-emphasis must be changed (R103, C103, R104) if other impedance microphones are used.

To summarize: adjust R2, R110, R105, and R106 so that each mode gives your transmitter about equal output power. Run slightly into ALC by adjusting R110 after the earlier adjustments. You will probably want to run the 2:1 compressor under normal circumstances where background noise is no problem.

Setting up the receive adapter is sometimes more critical. First, the gain of the audio power amplifier following the unit is undefined (I don't know what you plan to use or its drive and gain requirements — but I recommend a TA 7205 IC.)

This amplifier should be capable of several watts of output for best results and be able to give about 2 watts output with less than 0.25 volt p-p into 10 kilohms at its input. Most output stages will do this (that is, inserting audio into the adapter from the top of the potentiometer then reinserting after the potentiometer). Many hi-fi amplifiers require 1 volt or more rms or several volts p-p, however. The NE570 will not provide several volts. A buffer stage must be provided in these cases (perhaps made from the remaining stage of the 3403).

With the unit in the expand position, the volume from your receiver and to your audio amplifier/speaker should be adjusted to a level somewhere below the point at which the expander starts to cause distortion. This should be adjusted with a clean signal using fast AGC (or on a pre-compressed station) and with the receiver audio in its normal position. Rv can be used to set the gain through the expander to your audio stage so that it does not overload. Your amplifier's gain can then be adjusted for a desired listening volume. Resistor R101 can then be adjusted so that the audio is at the same level as the expanded audio when the expander is switched out.

Resistor R4, by the way, should be adjusted so that Rv can provide up to 10 dB gain at full volume or can reduce signals to levels well below the input level. Resistor R4 could even be replaced with a 4.7 kilohm resistor.

When all is adjusted properly, signals coming through the adapter will be roughly the same peak level with or without de-emphasis and/or expansion. However, expansion using the 600-Hz bandpass filter only may be slightly higher in volume than with the 800-Hz lowpass filter. Furthermore, when both filters are used on an unprocessed incoming signal, the volume may seem lower due to suppression of high and low frequencies. Some back-and-forth adjustments between Rv and your amplifier's volume may be required before you find the right combination for your system. Adjustment of the value of R101 may be needed also.

To summarize, the level to the NE570 must be set so it does not distort. The value of R101 may then be set to balance the in-out levels. The 600-Hz bandpass filter may show slight volume increase when used with the expander. The 800-Hz lowpass filter provides the best "punch" and should be used in circumstances when this is useful. Expand only should be used with fast AGC if the station received has no clipping and/or compression. The full system should yield significant SNR enhancement at 10-dB SNR or higher levels. Adjacent channel and high- or low-pitched carriers and interference should also be reduced due to both filtering and expansion.

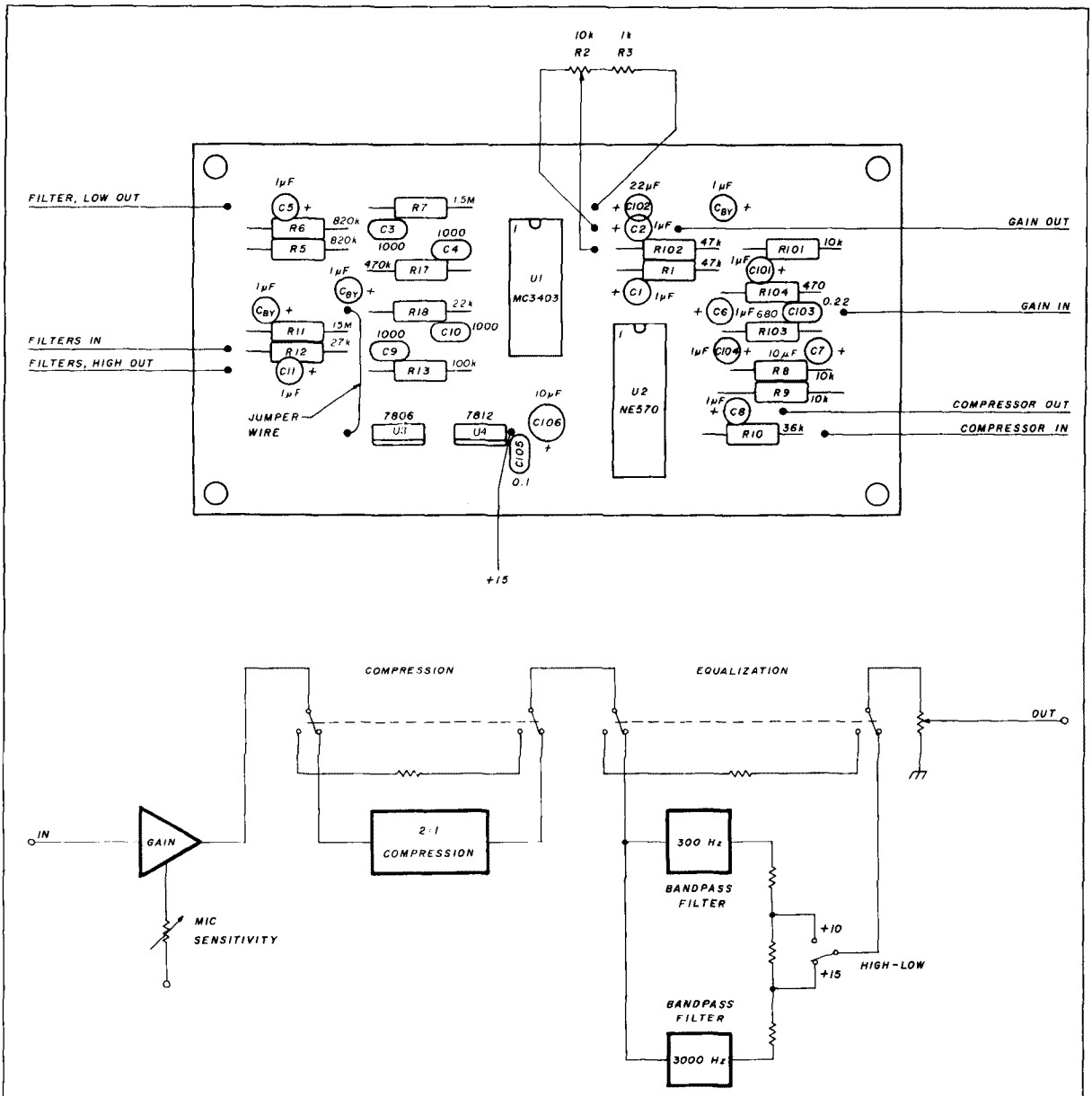


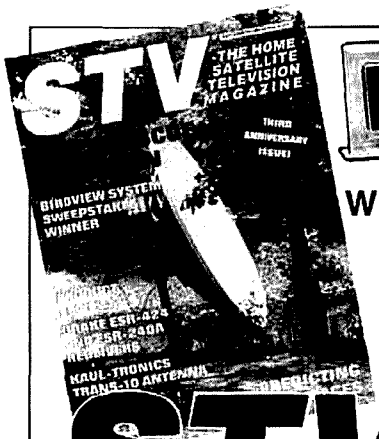
fig. 7. Level-one transmit adapter: (A) component placement diagram and (B) block diagram.

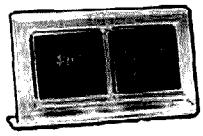
the receiver's volume control and the receiver's audio stage. This provides the best overall performance and is recommended.

The Radio Shack TA-7205 audio output IC is a good choice for a 3 to 5 watt output stage. It is readily available and comes with a schematic diagram. It can be built into a third Pomona Box or on a heat sink mounted in the external speaker box. I do not recommend such old standby ICs as the LM380, which does not have enough "headroom." Certain 2 watts per channel hi-fi parts with built-in heatsinks can be paralleled with good results, however.

Parts placement diagrams for the receive and transmit adapters are provided in **figs. 6** and **7**, respectively.** No special construction procedures are required beyond good RF shielding and grounding. When used with OSCAR stations, for example, it isn't acceptable

**An information sheet including prices on adapter PC boards is available from *ham radio*, Greenville, New Hampshire 03048. Please enclose SASE with your request.





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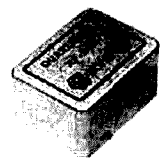
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to have audio rectification in the audio amplifier coming from your uplink transmitter! This is one reason we chose the Pomona 2901 box. This box, once the paint is scraped off around the internal lip and screw mounting posts, provides excellent shielding. Furthermore, it is easy to drill and looks very nice when the project is completed.

Radio Shack also sells several nice boxes for those who desire to put both transmitter and receiver adapters into one box. Be sure to use a few more screws than you normally would and scrape off a little paint here and there to ensure good RF shielding. Also remember that looping to both transmitter and receiver circuits from a common external box can create ground-loop problems.

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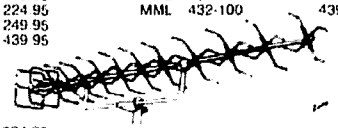
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"K7DLN, this is WB5AOX in Artesia. Has that storm reached Phoenix yet?" ...not an unusual ham transmission — except that WB5AOX is sitting in his kitchen, talking to a mobile station 500 miles away on a 2-meter HT. This kind of conversation has come to be perfectly routine in the area of the southwestern United States covered by the Zia Connection, a VHF/UHF communications super-system that takes its name from the ancient sun symbol of the Zuni Indians.

Zia links eleven 2-meter repeaters in what amounts to one 2-meter repeater with tremendous coverage. The dotted line in **fig. 1**, illustrating the area covered, represents the approximate limit of the system for an average mobile station running 10 to 25 watts with a 5/8 wavelength antenna. If the range available to fixed stations with large antennas or remote bases were included, the area covered would be even larger; as is, about 100,000 square miles are accessible to mobile stations within the system. About 1300 miles of six separate interstate highways are covered. Six hundred-mile QSOs between handheld stations are routine.

The major repeater sites include Tucson (population: 350,000); Phoenix (800,000); Albuquerque (350,000); El Paso (450,000); and Ciudad Juarez, Mexico (600,000). Given these numbers — and the nature of the geographical area covered — opportunities for public service (conducting weather watch and swap nets, reporting highway accidents, providing travel directions and locating travelers) abound. Hams are never out of touch in this vast area, even if they're miles from the nearest telephone . . . and much of the area *is* miles from the nearest telephone! In October, 1983, when flash floods wiped out telephone service to communities throughout the center of the region for several days, Zia provided trouble-free communication for the Red Cross and National Guard until services were restored.

Obviously Zia didn't spring into existence without precedent or without a lot of work by many hams. The western United States has long been noted for long-range repeaters situated on high mountain tops. Partly because the population density is low, there's been a drive toward even greater coverage. During the 1960s, shortly after 2-meter FM repeaters were first introduced to the region, there was a time when the major New Mexico repeaters were linked together. The first regular link between New Mexico and Arizona was established in 1977 between the 16-76 repeater of the Mesilla Valley Amateur Radio Club and the 31-91 repeater of the Eastern Arizona Amateur Radio Society. This was a part-time, tone-controlled link between well-established repeaters. Because the sponsors often differed as to how the link should be operated, however, it was eventually discontinued.

The Zia Connection acts just like a single wide-range repeater. The links are completely transparent to users and the whole system is open 24 hours a day to all who wish to use it, without any controlling necessary. The only time portions of the system aren't available is when technical difficulties or emergencies occur. There are no multiple squelch tails and no lost words at the beginning of a transmission. If one of the linked repeaters is timed out by a long-winded user, only that repeater — not the whole system — is shut down.

Each 2-meter repeater is similar to each of the others. All have CW identifiers and the same time-out timers, courtesy beeps, and controlled audio levels. The equipment must be highly reliable, available 24 hours a day, and able to withstand a high duty cycle. All stations are well designed, with such refinements as notch filters in the audio lines of the link receivers to remove the CW identification of the links, which are on 420 MHz.

Figure 2 shows the block diagram of a typical repeater node. The main component is the 2-meter repeater shown at the top. There is provision for up to three

By Alex F. Burr, K5XY, 2025 O'Donnell, Las Cruces, New Mexico 88001

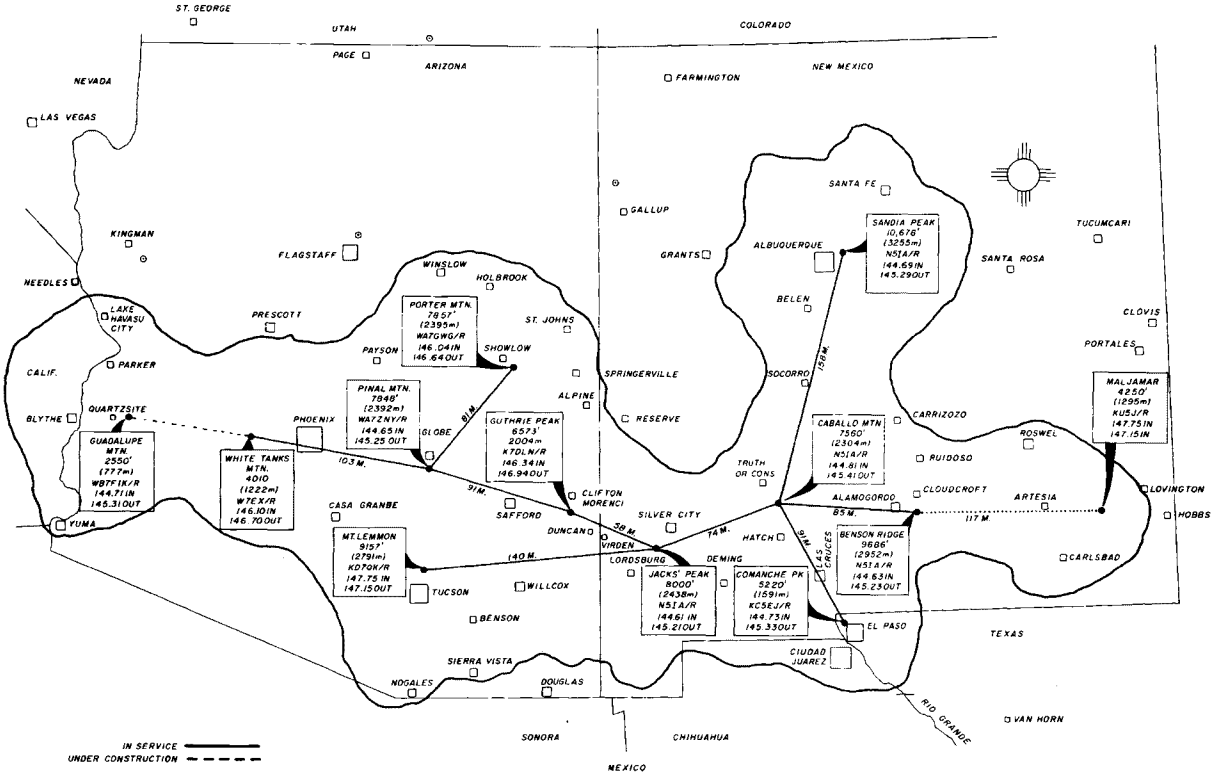


fig. 1. Mobile coverage range of the Zia Connection.

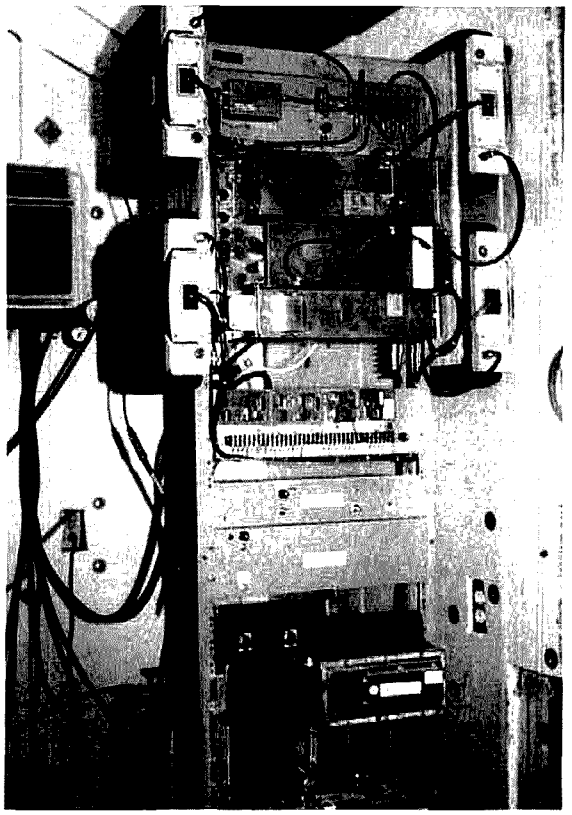


Photo A. The equipment rack at the Caballo Mountain, New Mexico location of the Zia Connection. Four converted commercial transceivers for the UHF links can be easily spotted on the two edges of the rack. The four printed circuit boards which make the essential transceiver and repeater interconnections are located in the middle.

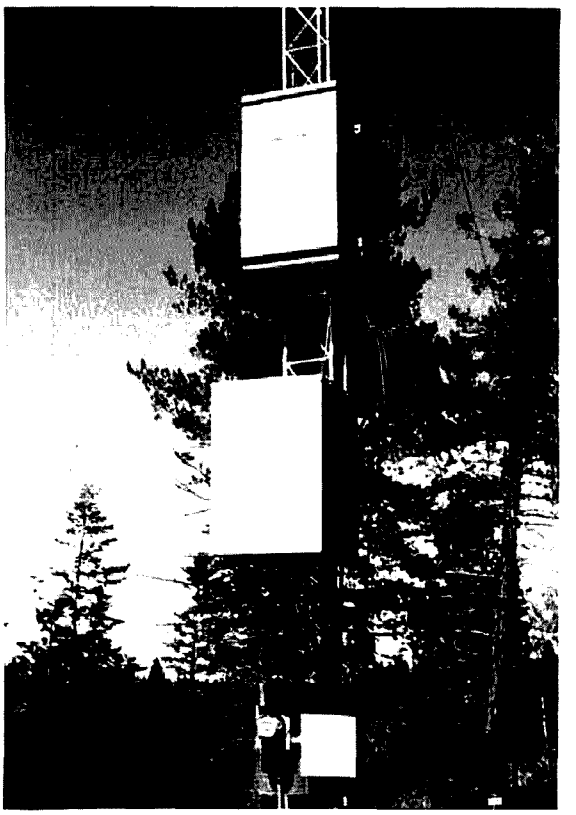


Photo B. Waterproof boxes near the base of the Zia Connection tower at Benson Ridge, New Mexico house all equipment.

UHF channels to act as control links. Two additional link ports are available with the addition of one more link board.

All three subsystems use typical Amateur equipment. The main antenna is a commercial grade 2-meter antenna; the UHF link antennas are home-brewed. The duplexers are standard, as are the various receivers and transmitters, which are mostly retuned commercial rigs. The links are full duplex, so there's only minimal delay in switching them.

The equipment used by the Caballo Mountain 144.81-145.41 repeater in central New Mexico, typical of what's available at each node, is a combination of commercial, modified commercial, and ham-built equipment. Four surplus Motorola MoTran 406-420 MHz transceivers used for the UHF links (**Photo A**) underwent minor modifications to insert audio level controls, bypass internal audio amplifiers, speed up the squelch action, and provide full duplex capability. The original 12-VDC power supplies have been replaced with external 24-volt power supplies. A General Electric MasterPro, modified in a manner similar to the UHF transceivers, is used for the 2-meter repeater. The 2-meter duplexer is a WACOM; the UHF duplexers are Phelps Dodge mobile duplexers.

The 2-meter feedline is Heliax and the UHF feedline uses RG213 coax. The antennas are mounted on a TV translator tower; the 2-meter antenna is a Phelps Dodge 340 and the UHF directional antennas — either 6- or 14-element verticals, depending on path requirements — are home-brewed models patterned after KLM designs. **Photos B** and **C** show another typical Zia installation at Benson Ridge, New Mexico.

Of course the key to the whole operation is the rather uncomplicated control system (**fig. 3**). With the emphasis on audio quality and fast keying response, it's designed to have a dropout time (time of response to loss of carrier on the 2-meter receiver) from one end of the system to the other of about 200 ms. The pickup time (time of response to the presence of a signal on one of the receivers) is about 300 ms for the complete system.

The control system consists of four main parts, three of which are quite conventional. Based on an EPROM,

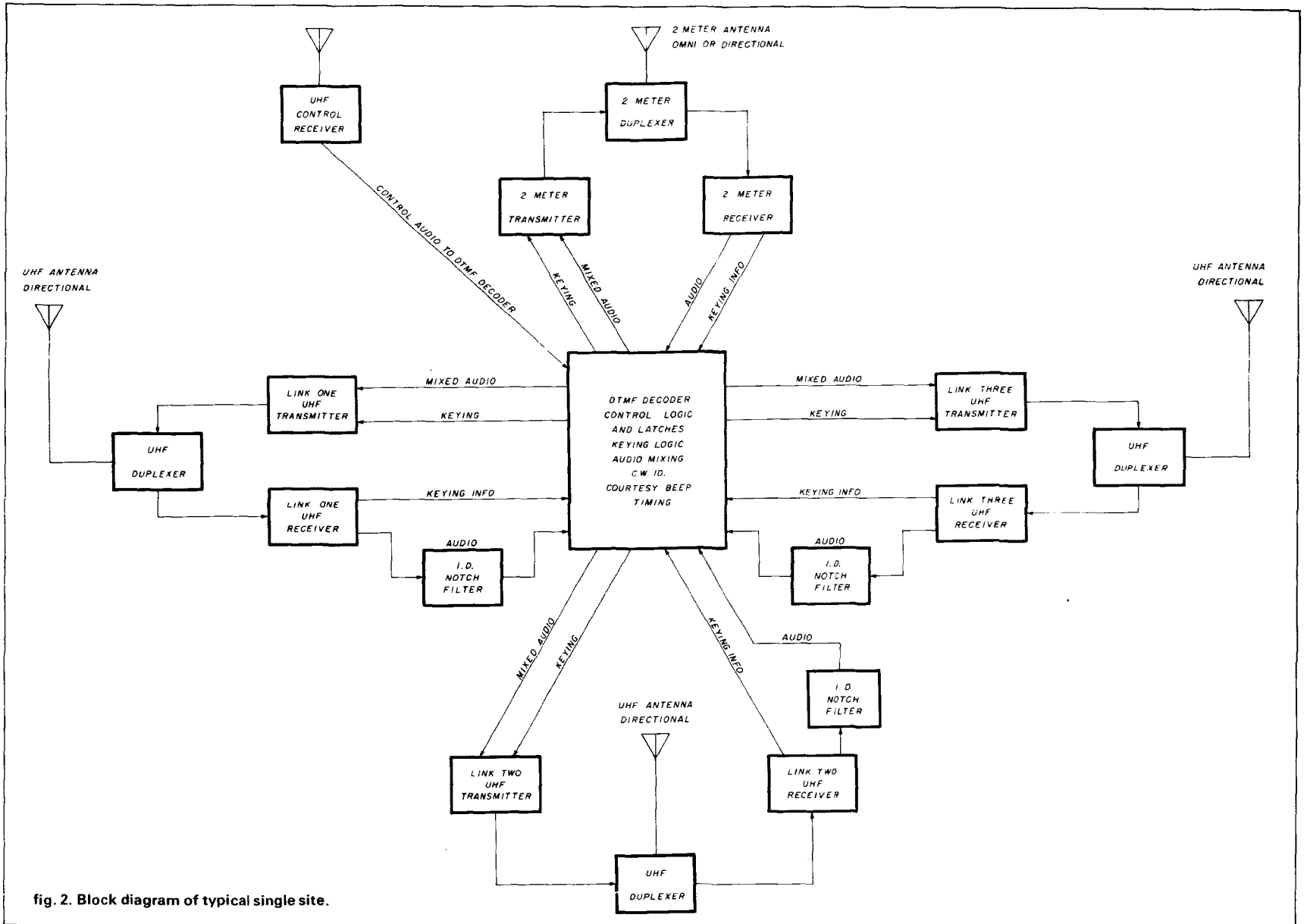


fig. 2. Block diagram of typical single site.

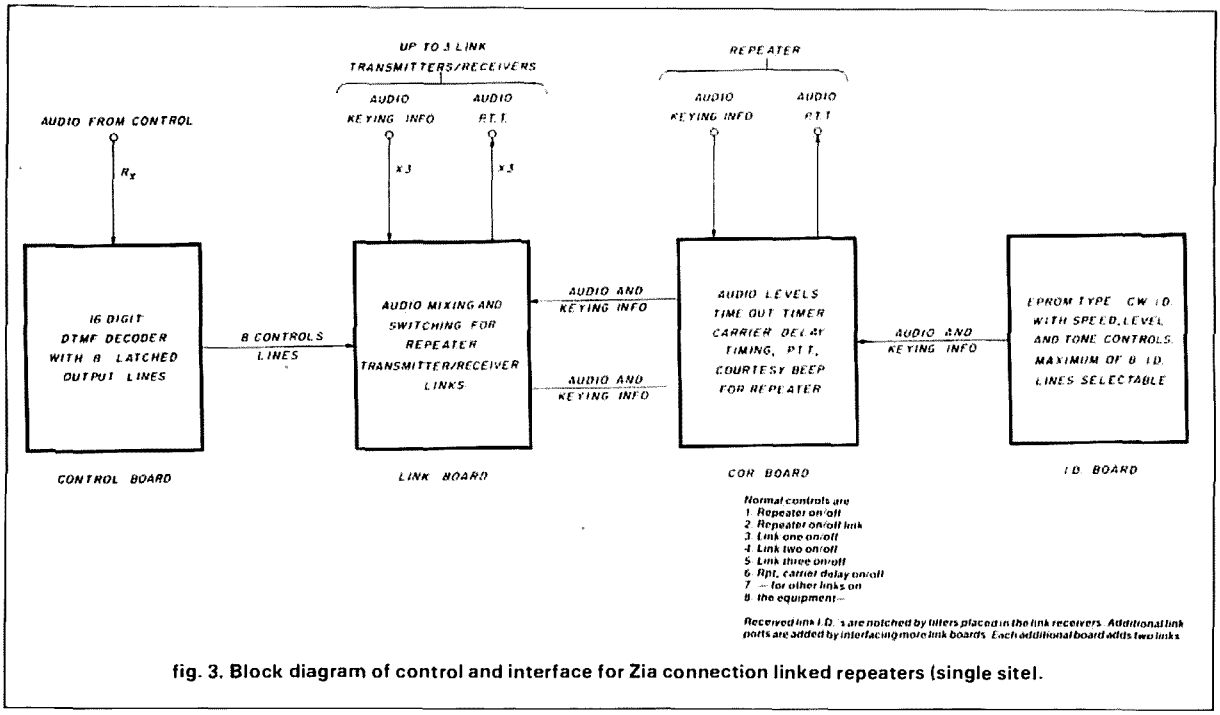


fig. 3. Block diagram of control and interface for Zia connection linked repeaters (single site).

the identification board (I.D.) on the right-hand side of the block diagram shown in **fig. 3** provides the required CW identification signal with speed, level, and tone controls. Its only unusual feature is the maximum of eight identification lines from which to select.

The control board on the left accepts audio from the UHF control receiver to a 16-digit DTMF decoder with eight latched output lines. Six of these lines are reserved for specific functions: the first turns the repeater on or off, the second connects the repeater to the link or lets it operate as an isolated repeater, the next three turn the three link systems on or off, and the sixth turns the repeater carrier delay on or off. The remaining two lines are for any other link equipment that might be added. All controls are independent of the others. The input to a particular board is accessed via UHF by the control stations for this particular part of the system.

The carrier operated relay (COR) board, also quite conventional, allows for setting audio levels, time-out protection, carrier delay timing, push-to-talk, and the courtesy beep for the repeater. The only unusual features are the audio and keying information input from the links, a similar output, and a mixing circuit to include the audio from the 2-meter receiver with the audio from the links.

The link board, the least conventional part of the system, has to process and mix the audio to and from as many as three links and the 2-meter repeater. It also has to provide and respond to the appropriate keying information and accept the eight output lines from the control board.

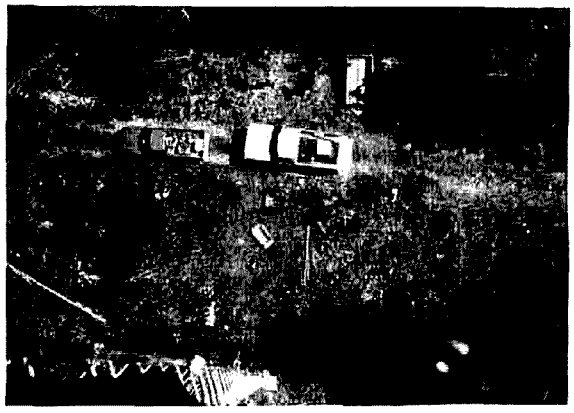
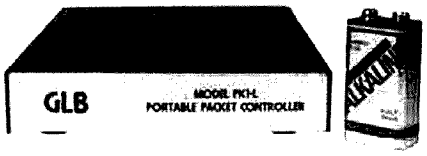


Photo C. View from the top the Zia Connection tower at Benson Ridge, New Mexico. The photograph was taken during construction.

It should be obvious that no single piece of the system's hardware is particularly sophisticated; instead, the Zia Connection is more of a triumph of organization and cooperation, of hard work and commitment, than of technology. Of the several factors in its successful operation, the most important is the dedicated leadership of Milt Jensen, N5IA, who started the network with three linked repeaters in August, 1982. Milt is the first to ad-

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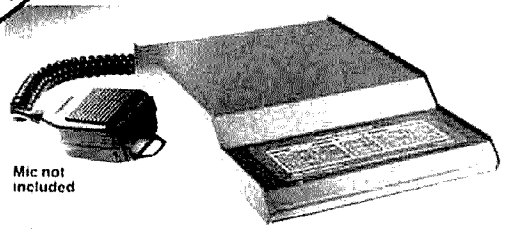
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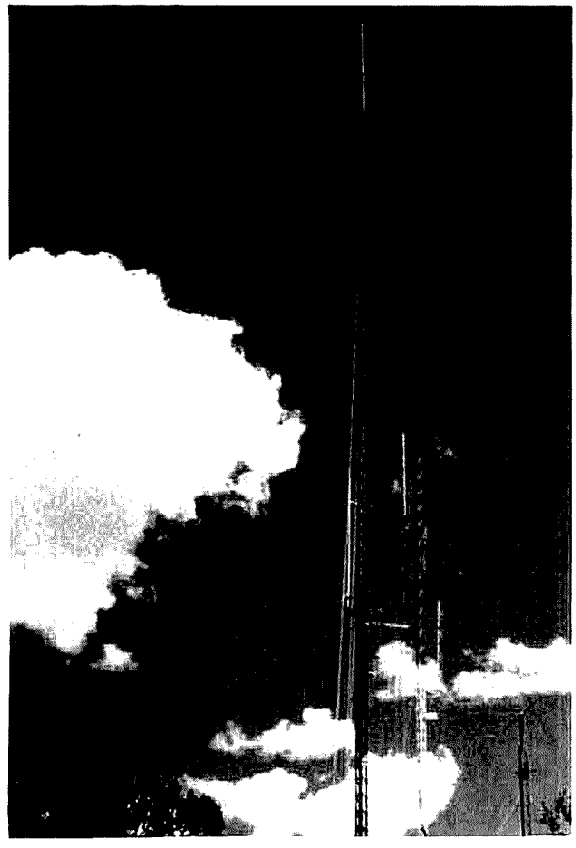


Photo D. The 135-foot tower used by the Zia Connection at Jack's Peak, New Mexico. The tower supports 11 antennas used by the 2-meter and UHF equipment. It took 12 people two and a half days to prepare the base, erect the tower, and mount the antennas.

mit that he hasn't done it alone, but has relied on the help of many others who've designed and built components, set up repeaters, made repairs, and performed regular maintenance. Site acquisition and frequency coordination have been no small jobs, either. But it was N5IA who decided to establish the Zia Connection, set the first parts in place and got them working, and continues to plan, guide, and help with expansion.

Because previous experience with repeater linking showed that it's difficult for established groups to agree on a single plan, N5IA decided to start afresh, without benefit of existing repeaters or radio clubs. Only individuals who would wholeheartedly support the basic concept were recruited. **Table 1** lists repeater sites and equipment owners; fortunately, at about the time this operation was begun, the number of repeater channels was increased so that new repeaters could be installed with little or no inconvenience to users of existing repeaters.

Milt received signficant assistance from Joe Montierth, WA7ZNY, who designed the control and inter-

Table 1. Zia connection nodes.

Site	Altitude (feet)	Equipment Owner
Mt. Lemmon	9157	John Vanza, AK7Z
Porter Mountain	7857	Kachina Radio Club
White Tanks	4010	Ken Simpson, WB7DRD
Jack's Peak	8000	Milt Jensen, N5IA
Caballo Mtn.	7560	Milt Jensen, N5IA
Benson Ridge	9686	Milt Jensen, N5IA
Sandia Peak	10,678	Milt Jensen, N5IA
Comanche Peak	5220	Rick Fultz, KC5EJ
Guthrie Peak	6573	Eastern Arizona Amateur Radio Society
Maljamar	4250	Bob Perkins, KU5J
Pinel Peak	7848	George Lewis, WA7SBZ, & Joe Montierth, WA7ZNY

face circuits and performed many of the equipment modifications. Rick Fultz, KC5EJ, provided the control and interface circuit board layouts, and Norm Smith, K7DLN, built the link antennas.

An example of the kind of cooperation common to Zia activities was the Jack's Peak antenna-raising party, coordinated by N5IA, which was necessitated by a severe ice storm that coated guy wires with 5 inches (i.e., 4 tons) of ice and brought the tower down (see **Photo D**). The operation involved building a concrete base (with 3 cubic yards of concrete mixed in a 1/3-sack mixer), raising a 135-foot tower, and setting eleven antennas and feedlines in two and a half days. All this was done on top of an 8000-foot mountain 150 miles from the nearest large city. Twelve people — including two from El Paso (200 miles away), two from Tucson (200 miles away), and two from Los Angeles (800 miles away) — were recruited for this project. The group even included a video tape camera operator who shot many minutes of tape showing the tower being slowly inched into position.

Several factors over which nobody has any control contribute to the success the Zia Connection. One is the low population density of the area. There are enough hams to ensure that *somebody* is likely to be listening, and therefore able to respond to a request for emergency aid, but not so many that the channel is always busy. The present duty cycle is a comfortable 65 percent. Another factor is the apparent dearth of "squirrels" (or "nuts," if you prefer) among the ham population of the region. The network hasn't been significantly bothered by unstable operators and everybody gets along without the need for formal rules or control operator intervention. No area of the system has preference over any other area; in all aspects, the system operates as if it were just one large, friendly repeater.

Of course the availability of mountain-top repeater sites helps, although remote mountain sites do have their disadvantages. But N5IA is quick to point out that the system can be duplicated, and even improved upon, anywhere that people have the desire to do so. A greater number of links, for example, could compensate for an insufficient supply of barely-accessible mountain-top sites.

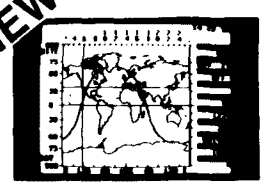
The separate units in the network belong to individual owners and two small clubs, who are responsible for all aspects of their repeaters and the associated links. Owners or owner-recruited volunteers such as Martin Raue, WB5LJO, Jim Devenport, WB5AOX, and John Braden, K7LKL, are responsible for maintenence. N5IA owns four of the eleven repeaters; others own the rest. Financing has been mostly by individuals, with some donations for maintenance and operation. There are no dues, and no soliciting; donations are made for the purpose of contributing to a good cause, rather than, for example, expecting *x* hours of operating time.

Even after the network was established and running, lots of effort had to be — and still must be — expended to keep it going. Time is the biggest problem; N5IA often feels that he has to be in several places at once. When you have to drive 150 miles (one way) to repair a repeater, it's bound to cut into your free time. Storms, with their antenna icing or lightning-induced failures, present the greatest challenge.

As it is, the region is fortunate to have a group of Amateurs who find pleasure in providing a remarkable service to a large number of hams.

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Bill Orr W6SAI

the grounded grid amplifier: part 2

In my last column I discussed the grounded grid (cathode driven) amplifier with respect to neutralization, RF feedback, flashover, and operating filament voltage. This column continues the discussion with respect to other features of this interesting circuit.

metering circuit

Correct operation of the grounded grid amplifier is determined by observing the grid and plate currents of the tube. Ideally, two meters are required; these should be placed in the circuit where they measure the wanted currents and no others, and where the meters are at (or near) ground potential so they present no high voltage hazard to the operator. Placing a meter in the B-plus lead, for example, is risky business unless the meter is specifically designed for operation at high potential above ground. Most meters are not designed for this type of operation.

Figure 1 shows one solution to the metering problem. Since all tube currents flow in the cathode (filament) circuit, one terminal of each meter is common to this circuit at point A. Note that chassis ground is taken as the grid of the tube, point B. In terms of DC, point A is therefore "above ground" by the voltage drop across the grid meter, M1. And since the plate current meter, M2, is in the negative lead of the high voltage supply, the negative terminal of this supply is "above ground" by the voltage drop across meters M1 and M2.

The difference in potential between the negative of the plate supply, point A, and chassis ground, B, is only a few tenths of a volt, but it's sufficient to allow metering to take place. Plate current flows from the filament (cathode) at point A, through the plate meter and plate supply to the anode of the tube. Grid current flows from point A, through the grid meter to the grid of the tube, which also happens to be point B, or chassis ground.

Although this circuit works, it isn't safe. Assume there is a DC arc to ground in the plate circuit. A typical arc in air has a resistance of about 12 ohms. This means that the whole energy in the power supply is "dumped" into the arc (**fig. 2**). Most Amateur transmitters have a high capacitance filter arrangement that stores many joules of energy. The arc current can be several hundred amperes for a fraction of a second. During the arc, the plate side of the supply is effectively at ground potential. This means that the negative side of the supply is now above ground by a large fraction of the supply voltage. At the same time, the high arc current flows through the plate and grid meters. *ZAP!* The meters literally disintegrate and the negative terminal of the supply is now above ground by the amount of the plate potential. *This is a truly deadly situation that could kill you if you were in contact with the equipment negative and chassis ground at the same time.*

The same can occur if the plate meter is overloaded and opens up. Again, the DC return to the power supply is opened and the negative terminal floats above ground at a dangerous potential.

The whole point is that dangerous conditions can exist when meters are simply placed in the negative return leads of a power supply without the inclusion of protective measures.

The circuit designer can protect the user from this lethal situation by never allowing the negative terminal of the power supply to leave near-ground potential. This can be accomplished by placing a low-value resistor, R, from the negative of the supply to chassis ground. The value of the resistor must be high enough so that it doesn't alter the accuracy of the meter readings to

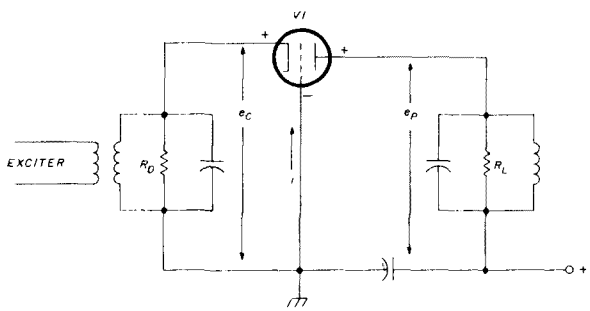


fig. 1. Grid and plate meters are placed in low-potential leads to protect the operator. Circuit creates as many problems as it solves.

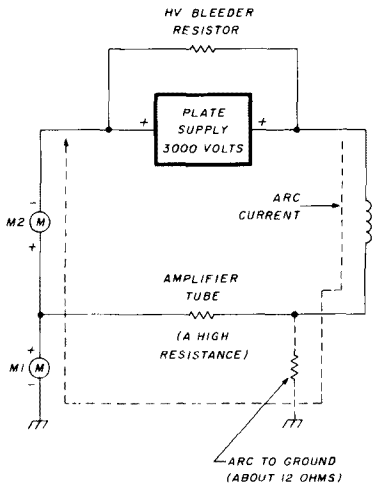


fig. 2. The plate circuit from a DC point of view during an arc.

any degree. Usually several hundred ohms will do the job. The resistor can be placed from negative to ground directly, or it can be split in two, with separate resistors placed across each meter. I prefer to use a single resistor to ground, as shown in **fig. 3.**

How big (physically) should the resistor be? That depends upon the energy storage capability of the power supply. Hams who blithely use hundreds of microfarads of filter capacitance in their supply (in the mistaken idea that it's required, and will make their signal "louder" in some magical way) are asking for trouble in case of a DC arc in the system. In most cases, not more than 20 μF is required, and sometimes less than that. Even so, it takes a large-wattage resistor to stand the gaff during a DC short situation. A small-wattage resistor will explode with a flash that sounds like a cannon going off in the room. I use a 200-ohm, 50-watt resistor in my supply; so far it has withstood a bad flashover in the DC portion of my plate circuit with no apparent damage.

protecting the meters

If there's a short or DC arc in the power circuit, the meter will pass the current until a fuse blows or the meter burns open. A meter fuse could be used — but if it were to blow, the circuit would be opened. The meter might be saved, but the circuit could float above ground and pose a hazard to the operator or perhaps to other components.

A good solution is to place reverse-connected diodes across the meters. Ordinary silicon diodes of the type used in power supplies have a 0.7-volt drop across them that's sufficient to allow the meter to function properly. Type 1N2071 or 1N4005 diodes will do the job. Even if the short current damages the diode, it commonly fails in a short circuit mode, protecting the rest of the circuit. **Figure 4** shows the protective system across representative meters.

vacuum-arc protector

It's possible for any power tube operating at a moderately high plate potential to experience a vacuum arc. This is known as the "Rocky Point" effect, named after the RCA transmitting site where the phenomenon was first noticed in the 1930s. The vacuum flashover may be brought about by a small number of gas ions in the tube, or even by bombardment of the tube by cosmic ray particles from outer space! If the transmitter is power supply-limited, the quick discharge has no damaging consequences. However, most ham transmitters have a high capacitance filter in the power supply whose energy is instantaneously discharged into a flashover. Extremely high currents can flow in a fraction of a second, damaging the tube or associated circuit components.

It's easy to protect the tube and equipment by placing a protective resistor, R1, in the B-plus lead to the tube (**fig. 4**). The vacuum arc is sensitive to voltage and the voltage drop across the resistor during an arc is enough to extinguish the arc. In most instances, the operator will be unaware of an arc taking place, since the whole action takes place in less than a millisecond.

complete protective circuit

The metering and protective circuit, with all components included, is shown in **fig. 4**. The additional components are inexpensive and can save plenty of more valuable components from biting the dust when trouble occurs in the amplifier.

filament voltmeter

All power tubes have a filament voltage tolerance specified by the manufacturer. In many cases it's ± 5 percent of the nominal value. For a tube with a 5-volt filament, the limits, therefore, are 4.75 and 5.25 volts.

The primary power circuit in the United States is standardized on 117 volts, ± 5 volts. That's about ± 4 percent. Variations in primary line voltage can thus "eat up" the filament voltage tolerance of the power tube.

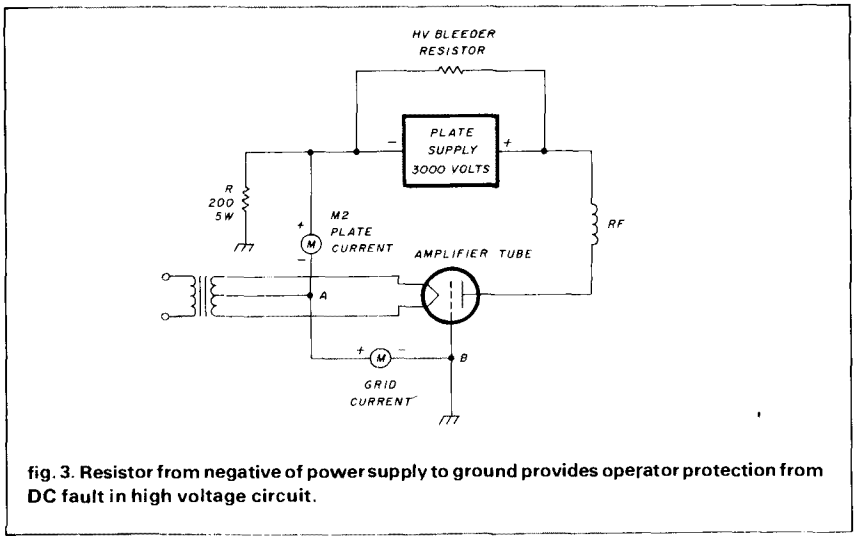
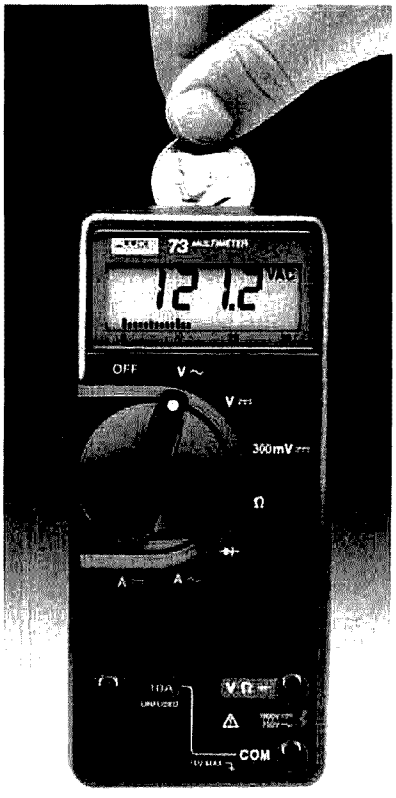


fig. 3. Resistor from negative of power supply to ground provides operator protection from DC fault in high voltage circuit.

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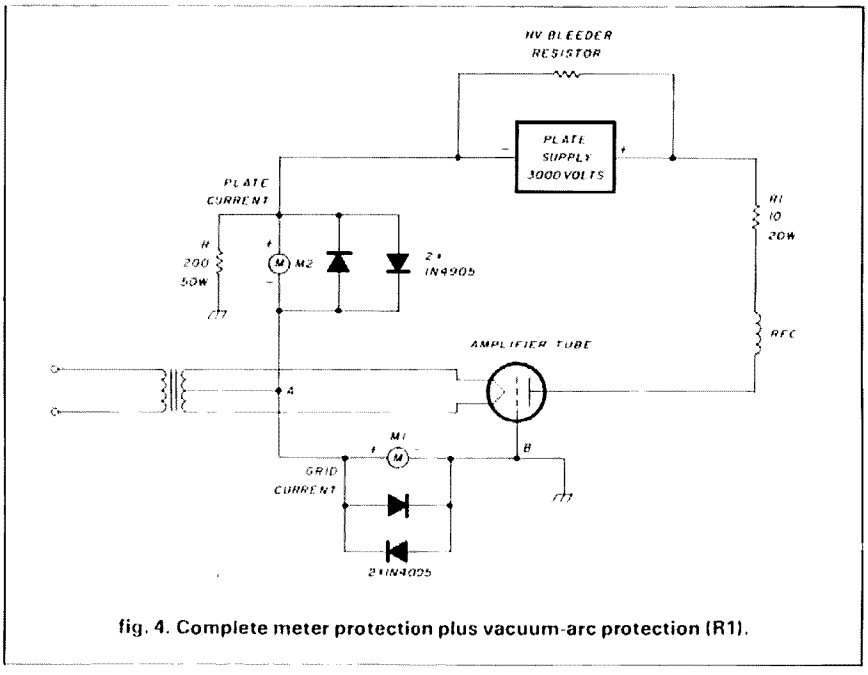


fig. 4. Complete meter protection plus vacuum-arc protection (R1).

Many prudent Amateurs run the filaments of their power tubes from a separate filament transformer set to the correct voltage by means of a variable voltage transformer ("Variac" or "Powerstat" are two trade names for such a device.) The filament voltage is set "on the nose" by means of a filament voltmeter. Or is it?

Primary voltage in most of the United States is affected by industrial loads that can alter the waveform from the classic sine wave. The primary line is full of harmonics, spikes, and other unpleasant surges and dips. This makes an easy job difficult because inexpensive volt-ohmmeters and other instruments that employ a rectifier and DC meter to measure AC voltage simply lose their accuracy when confronted with a distorted waveform.

An accurate method of measuring filament voltage is to use an "RMS responding" AC meter. The old Weston Model 301 AC instruments often found at flea markets are of this type. Beware of a meter bearing a label that says "rectifier-type instrument." It won't do the job as well.

When you find an RMS-responding AC meter, take it to an instrument laboratory, if there's one nearby, or perhaps a local college. Ask their standards lab to calibrate your meter for you. This is a cheap and easy way to obtain a 1 percent meter for a few dollars and just a little effort. You can then use the meter to monitor your filament voltage or your line voltage.

I picked up a nifty 0-150-volt Weston laboratory-type meter with a mirror-scale for a few dollars at a local swap meet. A local standards lab checked the meter out, and now I have a "standard" meter, good to 1-percent accuracy. I leave it permanently connected across the primary line to observe the daily fluctuations of line voltage. In my case, the voltage varies with the time of day, ranging from 118 to 124 volts! Since it's always above 117 volts, I have a multi-tap filament transformer connected as an autotransformer. I usually set the voltage to my transmitter at 115 volts, slightly under normal, in the hope that everything will run cooler and with less stress than if I operated it directly off the line. The voltmeter is atop the operating table and the tapped transformer is suspended under the tabletop, to the right of my equipment.

After these alterations, I might not be any louder on the air, but I am a lot safer and my equipment is protected from misfortune, which often occurs at just the wrong moment.

ham radio

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get on SSTV — with the C-64

Software approach/ uses BASIC and/ machine language routines

Although the Commodore 64 computer has proved to be a popular addition to many hamshacks, its full potential hasn't always been realized. The Commodore has excellent graphics capabilities that are easily accessible to a knowledgeable programmer — and these capabilities are directly applicable to slow-scan television.

Several years ago, Edwin Cox designed and published an SSTV send and receive program for the VIC-20, later upgrading it for the C-64.[1,2,3] Others have written programs to allow weather facsimile to be displayed using the Commodore machines.[4] Numerous articles and books have explored what can be done with the versatile Commodore machines[5]; the possibilities described in the Amateur literature go far beyond applications available commercially.

While the earlier programs required additional hardware in order to use the Commodore for visual reception, the programs described in this article make full use of the hardware already contained in the C-64. The only interface necessary is protection of the sensitive ICs from the fluctuations of the outside world.

What follows is the brainchild of two Italian computer and radio enthusiasts. Once you've typed in the programs, you'll be ready to send and receive black and white SSTV pictures with very little additional work. — Jim Grubbs, K9EI.

slow-scan basics

Before we program the computer for transmission and reception, let's briefly review the basics of SSTV. SSTV is transmitted as a continuous series of tones using normal voice bandwidth channels. While most SSTV activity occurs on HF using standard SSB techniques, it's equally well-suited to FM modulation on UHF/VHF or even transmission over standard grade telephone lines.

First the image to be transmitted is scanned. For a standard black and white picture, it takes about 8 seconds to scan and send a single frame of video. In order to know when a new scanning line begins and when a new picture frame begins, certain tones are reserved to indicate syncronization signals. A 1200-Hz tone represents a sync pulse. A horizontal sync pulse consists of a 1200-Hz tone for a period of 5 milliseconds. A longer 1200-Hz tone for 30 milliseconds represents a vertical sync signal.

In order to distinguish between black and white portions of the picture, the tones between 1500 and 2300 Hz are used to specify everything from pure black to pure white — including all the shades of gray in between.

making the computer speak SSTV

In order for the computer to interpret an SSTV signal accurately, it has to know exactly what frequency is present during each sampling period. By connecting the audio to the user port, (see **fig. 1**) of the Commodore 64, we can count the number of zero crossings present and determine the frequency of the incoming signals. We use pin "B" on the user port, which is available as a general-purpose interrupt input on the 6526 complex interface adapter chip.

The sampling of the input frequency must be done continuously using a very short time interval in order to reproduce a slow-scan picture accurately. This is achieved by rewriting the Non-Maskable Interrupt (NMI) routine that's resident in the C-64. By modifying the pointers to the NMI routine, we can substitute a program designed to suit our needs for SSTV. The pointers are located at hex locations $0318 and $0319.

taking a break

The central processing unit in the C-64 is kept very busy; all the pretty graphics and sounds available take their toll on processor time. For the purpose of SSTV reception, this can present a problem if we don't focus the attention of the CPU on our assigned task. Normal-

By Giuseppe Cameroni, I2CAB, and Giancarla Morellato, I2AED, Via Damiano Chiesa 26, Vigevano 27029, Italy. Translation and additional text by **Jim Grubbs, K9EI.**

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Table 1. SSTV reception program listing.

```
1000 rem %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
1010 rem direct reception of sstv
1020 rem pictures via commodore 64
1030 rem written by:
1040 rem giancarlo morellato i2aed
1050 rem address: via d. chiesa 26
1060 rem          27029 vigevano
1070 rem          italy
1080 rem %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
1230 print chr$(147)
1270 rem data loading
1280 :
1290 for i=32768 to 33285:read a:poke i,a:next
1300 poke 53280,0:poke 53281,0
1310 print chr$(147)
1320 print"sstv reception active"
1330 for i=1 to 5000:next
1340 sys 32768
1350 :
1360 data 120,160,000,132,158,169,096,133,159,169
1370 data 000,145,158,200,208,251,230,159,166,159
1380 data 224,128,144,243,160,000,132,158,169,004
1390 data 133,159,169,102,145,158,200,208,251,230
1400 data 159,166,159,224,008,144,243,160,000,132
1410 data 158,169,216,133,159,169,001,145,158,200
1420 data 208,251,230,159,166,159,224,220,144,243
1430 data 088,160,000,132,158,169,096,133,159,141
1440 data 033,208,169,080,133,159,169,188,145,158
1450 data 200,208,251,230,159,166,159,224,095,144
1460 data 243,169,241,141,134,002,173,000,221,073
1470 data 001,141,000,221,173,017,208,073,032,141
1480 data 017,208,173,022,208,073,016,141,022,208
1490 data 169,072,141,024,208,169,036,141,024,003
1500 data 169,129,141,025,003,169,127,141,013,221
1510 data 169,101,133,159,169,080,133,158,169,152
1520 data 141,015,221,141,006,221,169,003,141,007
1530 data 221,169,000,133,003,168,145,158,169,255
1540 data 141,004,221,141,005,221,173,013,221,169
1550 data 153,141,014,221,169,147,141,013,221,088
1560 data 032,228,255,240,251,201,136,208,011,173
1570 data 017,208,073,016,141,017,208,076,199,168
1580 data 072,173,000,221,073,001,141,000,221,173
1590 data 017,208,073,032,141,017,208,173,022,208
1600 data 073,016,141,022,208,169,021,141,024,208
1610 data 169,001,141,134,002,169,071,141,024,003
1620 data 169,254,141,025,003,169,142,032,210,255
1630 data 173,017,208,009,016,141,017,208,169,065
1640 data 032,210,255,088,104,104,104,104,234,234
1650 data 120,096,072,138,072,152,072,172,013,221
1660 data 169,127,141,013,221,186,224,064,176,006
1670 data 162,250,154,076,178,128,152,041,129,201
1680 data 129,240,243,152,041,145,201,144,208,039
1690 data 162,008,142,015,221,173,007,221,133,077
1700 data 173,006,221,070,077,106,070,077,106,174
1710 data 005,221,134,077,162,029,142,015,221,162
1720 data 003,221,246,129,176,013,202,016,246,230
1730 data 003,169,025,141,014,221,076,235,129,169
1740 data 077,073,255,074,133,077,074,024,101,077
1750 data 134,077,072,041,248,168,104,106,041,003
1760 data 073,003,170,165,077,024,202,048,004,010
1770 data 010,144,249,072,166,003,224,016,176,035
1780 data 224,004,144,048,230,158,208,004,230,159
1790 data 048,023,165,158,041,007,208,025,165,158
1800 data 024,105,056,133,158,165,159,105,001,133
1810 data 159,201,128,144,008,162,101,134,159,162
1820 data 080,134,158,169,000,132,077,168,145,158
1830 data 164,077,104,192,184,176,004,017,158,145
1840 data 158,152,105,008,168,169,000,133,003,145
1850 data 158,169,147,141,013,221,104,168,104,170
1860 data 104,064,052,085,106,123,000,075,160,105
1870 data 105,095,078,085,073,074,081,032
ready.
```

Table 2. SSTV transmission program listing (BASIC).

```
1000 rem %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
1010 rem   sstv transmission via c-64
1020 rem   conceived by:
1030 rem   giuseppe cameroni i2cab
1035 rem   and
1040 rem   giancarlo morellaton i2aed
1050 rem
1060 rem %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
1100 for i=19968 to 20079:read a:poke i,a:next
1110 for i=20224 to 20274:read a:poke i,a:next
1120 for i=20480 to 20696:read a:poke i,a:next
1130 for i=20736 to 20856:read a:poke i,a:next
```

(continued on page 47)

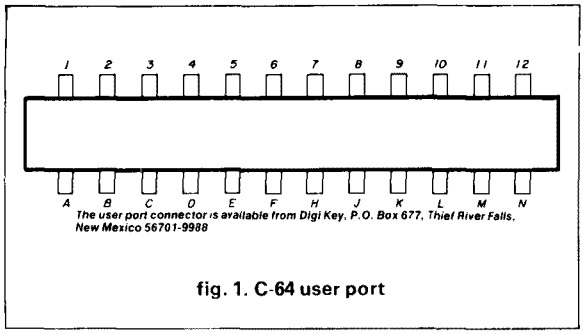


fig. 1. C-64 user port

ly the C-64 stops everything it's doing and takes time out to update the graphics screen. If these video interrupts are left intact, our program will work, but there will be "holes" in the picture where the CPU goes and does other things.

By making a small compromise in operating convenience, this problem is easy to overcome. The SSTV receive program has been written so that by pressing the F7 function key, the video interrupts normally processed by the CPU are suspended until the F7 key is pressed again. This results in the screen being blanked during reception of a picture, but ensures that no data is lost during the reception. (The Cox SSTV program suffered from the same problem and used a similar solution).

Because no external processing is being done, it's necessary to keep the audio gain of the receiver fairly high in order to saturate or clip the signal. The computer doesn't understand much other than square waves.

Two interfaces are presented here (see **figs. 2** and **3**). The first is definitely intended for those in a hurry, but it works! Just about any NPN transistor will do; its primary purpose is to protect the 6526 chip in the computer from damage. The second, a slightly more complex single-chip interface, allows a much more acceptable audio level to be used from the receiver. It uses a single LM311 voltage comparator IC and a few resistors.

getting the picture

Once you've typed in the program and built either of the two interfaces, you're ready to receive black and white SSTV pictures.

For convenience, the program for SSTV reception is presented in BASIC (**table 1**). Because the timing for SSTV is in the millisecond range, only a machine language routine is fast enough to provide the needed accuracy. This program "pokes" in the machine language program from BASIC.

To begin, connect the interface, making sure that the power to the computer is turned *off*; trying to connect the interface with power on could damage the C-64's sensitive circuitry.

Next, load and run the program. There will be a short delay while the machine language routines are poked into memory. After that the screen should turn black.

The best place to look for SSTV activity is around 14.230 MHz, particularly on Saturday and Sunday mornings. Be aware that several formats are in use; much of the activity you'll hear will be color transmissions not well-suited to this simple program. Many SSTV nets, however, ask their participants who work in color to also send their pictures in standard 8-second black and white format for those with less sophisticated equipment.

The best way to tune an SSTV signal is to adjust the receiver for normal-sounding speech and then leave the tuning alone during picture transmission. If everything is working smoothly, you'll see a picture begin to be traced in the center of the computer screen. The image will occupy only a portion of the screen.

The best images for this simple program will consist of graphics that don't contain subtle color changes. Signals generated using a SSTV keyboard or employing block-type graphics will come through best.

Using this program we've been able to display live images of distant SSTV operators. The display is somewhat grainy, but the image is still quite interpretable as a person.

Once you can receive pictures, you may want to transmit some yourself. While the simple program shown in **table 2** won't allow you to create and send fancy pic-

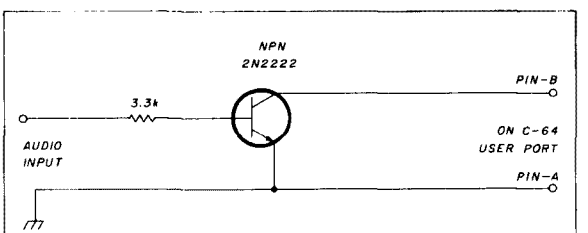


fig. 2. Simple receive interface

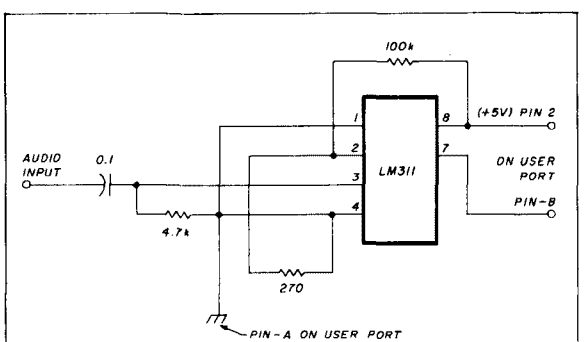


fig. 3. SSTV receive interface uses an LM311 voltage comparator.

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tures, it will enable you to call CQ and acknowledge other SSTV signals with one of your own.

SID sends a picture

Normally we think of the SID sound chip in the Commodore machine as being the hardware necessary to make music. For SSTV, however, it becomes an important part of getting a "picture" on the air.

The simple SSTV transmit program takes a portion of the regular text screen available on the C-64 and provides a conversion for everything typed in this special "window." The window consists of seven lines of up to eight characters each. Any of the standard Commodore characters, including the graphics characters, can be used. [I did find that the program acted strangely when I tried to send reverse-image characters. — K9EI]

The characters you type on the screen are presented, one at a time, to a special interpreter program. Every character consists of individual points of light; by processing these points and applying them to the sound chip, we can create an SSTV signal in standard format.

Since we're using only standard characters, all of the information will represent either pure black or pure white, depending on whether the point being "scanned" by the program is on or off. The computer itself will allow us to specify up to 255 different shades of gray. A more advanced program could use the main subroutines from this one to send a true bit-mapped image, perhaps prepared with one of the graphics programs like Koalapainter.

Every line of the video from our SSTV transmit program consists of 64 memory locations. A $00 specifies white, which is represented by a 2300-Hz tone sent for 0.93 milliseconds. A value of $FD indicates black, sent at 1500 Hz, while $FE indicates a horizontal sync pulse and $FF is used to indicate vertical sync.

Each frame of transmitted video consists of seven text lines containing eight characters each. Each character is represented by an 8 x 8 matrix. The image definition will be 63 points horizontal (8 x 8 less one horizontal sync pulse) by 128 points (7 characters by 8 x 2 — each line is transmitted twice — plus 16 lines of buffer). The extra lines of buffer were inserted to allow for easy reception of the picture even if the monitor isn't set up perfectly.

Finally the SID chip is turned on and the frequency is changed according to the values stored in the matrix. The resulting standard SSTV audio signal can be heard coming out of the monitor speaker.

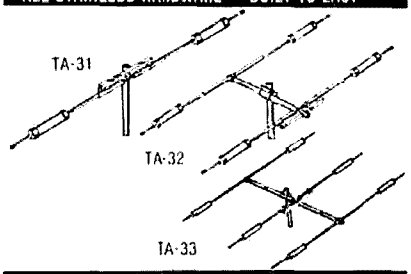
Once you've created your picture, you can send it by pressing the "backarrow" key located in the upper left hand corner of the keyboard. You can move the text window around using the cursor keys. You can return to the upper left hand corner of the window by pressing the RETURN key. To terminate transmission, you must press and hold the RETURN key. The key is scanned during the vertical sync period.

```
1150 print chr$(147):printtab(10)"sstv transmission"
1155 for z=1 to 8
1160 print chr$(17)
1170 next z
1190 printtab(3)"activate transmission by the ";
1195 print chr$(18)+chr$(95)+" key"+chr$(146)
1200 printtab(2)"stop transmission by pressing ";
1210 print chr$(18)+"return"+chr$(146)
1250 print chr$(19);:for i=1to8:print chr$(175);:next
1260 print chr$(19):for i=1to 7:printtab(8)chr$(212):next
1270 for i=1to8:print chr$(197);:next
1280 print chr$(19)
1290 for c=1 to 7
1300 for i=1to8:print " ";:next i:print
1310 next c
1330 print chr$(19):n=0
1340 p1=1064
1350 poke p1+54272,14:pokep1,peek(p1)+128
1360 get a$:if a$=""thenpokep1,peek(p1)-128:goto1350
1370 if asc(a$)=157thenprint a$;:poke p1,peek(p1)-128:n=n-1:p1=p1-1:goto 1350
1380 if asc(a$)=29 then print a$;:poke p1,peek(p1)-128:n=n+1:p1=p1+1:goto1350
1390 if asc(a$)=145then print a$;:poke p1,peek(p1)-128:n=n-8:p1=p1-40:goto1350
1400 if asc(a$)=17then print a$;:poke p1,peek(p1)-128:n=n+8:p1=p1+40:goto 1350
1410 if asc(a$)=95then pokep1,peek(p1)-128:goto1540
1420 if asc(a$)=13 then n=0:pokep1,peek(p1)-128:goto1330
1430 poke p1,peek(p1)-128:print a$;:p1=p1+1
1440 n=n+1
1450 if n=8then=p1+32:print
1460 if n=16 then p1=p1+32:print
1470 if n=24 then p1=p1+32:print
1480 if n=32 then p1=p1+32:print
1490 if n=40 then p1=p1+32:print
1500 if n=48 then p1=p1+32:print
1510 if n=56 then n=0:goto1330
1520 goto 1350
1540 ct=0
1550 sys 20224
1560 for i=1064 to 1351 step 40
1570 for c=i to i+7
1580 poke 2047+c,ct
1590 ct=peek(c)
1600 nn=0
1610 poke 56334,peek(56334)and254
1620 poke1,peek(1)and251
1630 po=1e+8+53248
1640 for p=potopo+7
1650 poke16384+un,peek(p)
1660 un=un+1
1670 nextp
1680 poke1,peek(1)or4
1690 poke56334,peek(56334)or1
1700 sys20736
1710 ct=ct+1
1720 if ct=56 then 1750
1730 nextc
1740 nexti
1750 for i=24576 to 32896 step 64
1760 pokei,254
1770 next
1780 pokei-64,255
1790 sys20480
1800 poke 54296,0
1810 goto 1330
1820 data 000,098,008,098,016,098,024,098,032,098
1830 data 040,098,048,098,056,098,000,102,008,102
1840 data 016,102,024,102,032,102,040,102,048,102
1850 data 056,102,000,106,008,106,016,106,024,106
1860 data 032,106,040,106,048,106,056,106,000,110
1870 data 008,110,016,110,024,110,032,110,040,110
1880 data 048,110,056,110,000,114,008,114,016,114
1890 data 024,114,032,114,040,114,048,114,056,114
1900 data 000,118,008,118,016,118,024,118,032,118
1910 data 040,118,048,118,056,118,000,122,008,122
1920 data 016,122,024,122,032,122,040,122,048,122,056,122
1930 data 169,090,141,016,079,169,126,141,036,079
1940 data 162,000,169,254,157,000,096,232,208,250
1950 data 238,016,079,169,098,205,016,079,208,238
1960 data 162,000,169,253,157,000,128,232,208,250
1970 data 238,036,079,169,129,205,036,079,208,238,096
1980 data 169,015,141,024,212,169,240,141,006,212
1990 data 169,017,141,004,212,169,127,141,013,220
2000 data 169,239,045,017,208,141,017,208,173,128
2010 data 128,201,000,240,029,201,253,240,057,201
2020 data 254,240,085,201,255,240,113,169,069,032
2030 data 210,255,096,234,234,234,234,234,234,234
2040 data 234,234,234,234,169,240,141,000,212,169
2050 data 152,141,001,212,162,001,160,153,136,208
2060 data 253,202,208,248,238,029,080,208,195,238
2070 data 030,080,076,028,080,255,169,000,141,000
2080 data 212,169,095,141,001,212,162,001,160,153
2090 data 136,208,253,202,208,248,238,029,080,208
2100 data 163,238,030,080,076,028,080,255,169,200
2110 data 141,000,212,169,079,141,001,212,162,000
2120 data 160,240,136,208,253,202,208,248,238,029
2130 data 080,208,131,238,030,080,076,028,080,000
2140 data 169,200,141,000,212,169,079,141,001,212
2150 data 162,097,160,130,136,208,253,202,208,248
```

(continued on page 51)

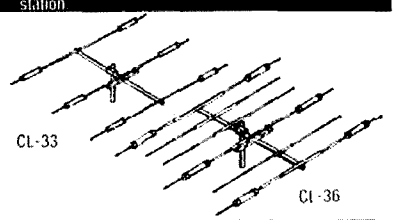
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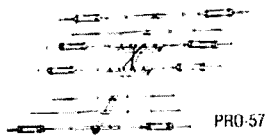


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(continued from page 47)

```
2160 data 169,129,141,013,220,169,016,013,017,208
2170 data 141,017,208,032,228,255,201,000,240,004
2180 data 096,234,234,234,169,000,141,029,080,169
2190 data 096,141,030,080,076,015,080
2200 data 173,255,079,010,234,234,170,189,000,078
2210 data 141,049,081,141,081,081,232,189,000,078
2220 data 141,050,081,141,082,081,162,000,160,000
2230 data 189,000,064,141,255,063,062,000,064,144
2240 data 005,169,000,076,046,081,169,253,153,184
2250 data 125,200,234,234,234,234,192,008,144,232
2260 data 160,064,173,255,063,157,000,064,062,000
2270 data 064,144,005,169,000,076,080,081,169,253
2280 data 153,184,125,200,234,234,234,234,192,072
2290 data 144,232,232,224,008,208,001,096,169,128
2300 data 024,109,049,081,141,049,081,141,081,081
2310 data 144,172,238,050,081,238,082,081,076,028,081
ready.
```

Incidentally, it does take a few seconds for the program to assemble the text into the matrix, so anticipate a few seconds delay when you press the transmit key. Like the receive program, the transmit program is presented in BASIC language form, although it does use machine language subroutines that are accessed from the BASIC program using the SYS command.

The SSTV audio tones will be present on the regular audio output connection of your monitor cord. If you're using a TV set as a monitor, the tones will appear in the TV speaker.

It's best to establish a direct connection between the audio output of the computer and your transmitter. You may wish to insert a small potentiometer in the line to allow adjustment for the proper level.

apples and oranges

How does this software-only method of SSTV reception and transmission compare with other methods that involve some external hardware? The original program published by Ed Cox in *vicComm* allowed SSTV reception on an unexpanded VIC-20 but required a two-chip interface. Because of the memory limitations of the VIC-20, the picture was a somewhat elongated, compressed image. The reception program presented here maintains the proper aspect ratios but doesn't produce a larger image. Later versions of the Cox program for both the VIC-20 with additional memory and the C-64 did give a full-screen presentation and at least got the aspect ratio pretty close.

On reception I found the Cox program and this one to be nearly identical in their level of clarity. They certainly aren't commercial quality, but are still a great way to get your feet wet on SSTV.

For transmit, the Cox program is somewhat more nicely configured, but does require the external interface. The picture quality coming from the software-only program is virtually identical, with the advantage of no external interface.

When compared to the program and hardware combination available from Kinney Software,* both programs show their limitations. While the Kinney approach is still not up to the state-of-the-art, it does produce a fairly high resolution image both in receive and transmit that neither of the other programs can produce. Still, it's a larger investment, so better results are to be anticipated.

While the SSTV software here has inherent limitations, it certainly shows what can be done by talented programmers with a machine as "simple" as the Commodore 64. With very little more than an expenditure of time, you can put your C-64 to work sending and receiving SSTV signals. Imagine what else it can do!

postscript

Questions or comments should be addressed directly to the authors at the address given at the beginning of this article, which was originally published in a somewhat different form in the national magazine of The Associazione Radioamatori Italiani (ARI).

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*Kinney Software, 121 North Hampton Road, Donnelsville, Ohio 45319.

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convert an inexpensive CB mag-mount antenna into a superb 2-meter whip

In mounting mobile antennas, position is everything

Radio Shack sells an inexpensive CB mag-mount whip antenna (Model No. 21-1005A, shown in **fig. 1**) that can easily be converted to hamband use. It's a rugged unit with a chrome-plated disc-shaped base housing the magnet and a 16-foot length of RG-58/U cable attached. The necessary modifications don't alter the external appearance of the antenna in any way.

From the tip to the bottom of the conical aluminum base, it measures 37-1/2 inches — close to the 1/2-wave resonant length at 146 MHz. It's tuned at CB frequencies by placing an inductor between the base of the whip and ground that resonates with the short monopole capacitive reactance. Tapped a short distance up the coil from the grounded end, it produces a 50-ohm resistive impedance.

modifying the base

Dismantling the base is easy. The conical aluminum base unscrews by hand from the black plastic housing, separating the aluminum base from the disc-shaped magnetic mount below. The plastic housing is then pulled gently from the magnetic mount, revealing the matching assembly: a tapped-inductor tuner! The only damage to the assembly caused by taking it apart is to a ring of caulking compound (probably silicone rubber) that breaks loose from the base of the black plastic housing. Weatherproofing can be restored — even improved — by replacing this and adding more caulking compound to the periphery of the hole where the RG-58/U cable emerges from the plastic housing.

matching network design

The existing matching system for the Archer antenna is ideally suited for modification into an L-network. The coil form, measuring 5/8 inch in diameter, is made of solid plastic. The No. 18 wire of the existing coil is entirely satisfactory. Even the mounting terminals on the coil form can all be conveniently used. A small variable capacitor is the only component that must be added. I used a compression mica capacitor from my junk box (the screw-adjustable kind, mounted on a little white rectangular ceramic base). Since the capacitance required will be in the order of 5 pF, I recommend one that's ad-

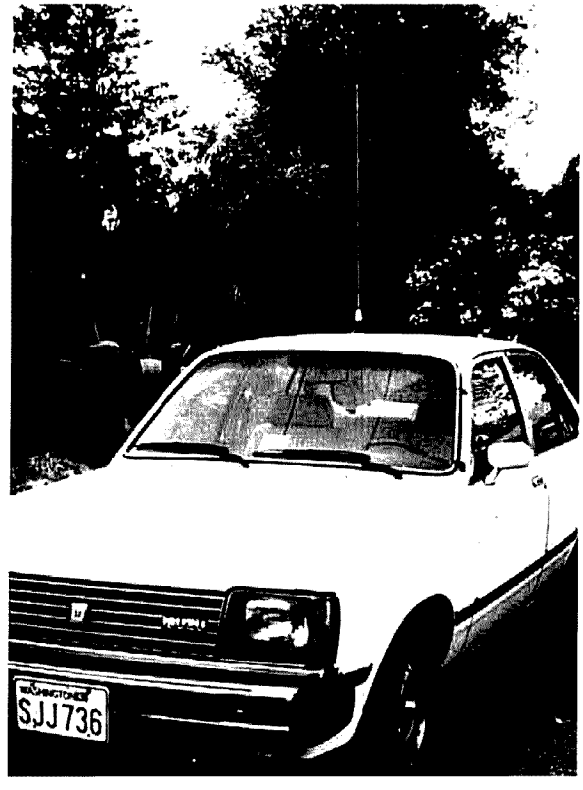


fig. 1. Mag-mount whip antenna after modification and installation.

By Donald K. Reynolds, K7DBA, 749 San Jude Avenue, Palo Alto, California 94306

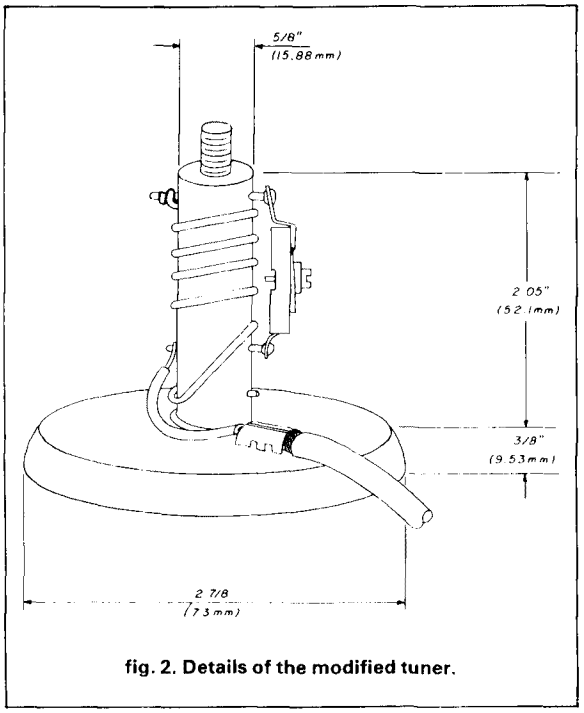


fig. 2. Details of the modified tuner.

justable in the 3- to 10-pF range. The ceramic-based compression mica type should handle powers of up to 100 watts.

I achieved a practically perfect match at 146 MHz by using the usual cut-and-try methods. My final coil had exactly five turns, from a bottom pin on the coil form (which is internally grounded to the magnetic base) to the pin directly above it at the top of the coil form, which is internally connected to the 1/4-20 screw onto which the aluminum base of the antenna is threaded. Since the spacing between the bottom and top pins to which the coil is attached is slightly over 1-1/2 inches, the turns are considerably spread out. In my final version, the top four turns are spaced about 1/4 inch apart; the bottom turn descends more steeply to reach the bottom pin. Three pins are on the coil form, each one extending horizontally through the plastic form so as to be accessible from either side.

The finished L-network is shown in **fig. 2**. The terminals on the capacitor have been bent out flat so that they can be soldered directly to the pins on the coil form. One terminal of the capacitor is soldered to the other end of the pin to which the center conductor of the coaxial cable is already soldered, and the other terminal of the capacitor is soldered to the opposite end of the pin to which the top of the inductor is soldered (that is, the base of the antenna). Nothing has to be done to the outer braid of the RG-58/U cable because it's already securely grounded to a lug under the coil form.

Figure 3 shows the schematic diagram of the final L-network. For those unfamiliar with the operation of these simple matching sections, they are ideally suited for transforming primarily resistive impedances from one value to another. Resistance *R*, the input resistance presented to the base of the whip and ground at the whip resonant frequency, will have a value in the neighborhood of 1000 ohms. The load resistance is shunted by the inductor so that the parallel combination is equivalent to a resistance of 50 ohms, in series with an inductive reactance. This reactance is then tuned out by the capacitive reactance of the adjustable capacitor. Assuming that the antenna represents a pure resistance of 1000 ohms at 146 MHz, calculated values for the L network are: L = 0.25 μH and C = 5.0 pF.

measured results

The final tuning of the antenna should be done with the antenna mounted at the selected location on the vehicle. It can be done with the black plastic covering of the matching network removed and the whip screwed to the top of the coil form. With a reflected power meter in series with the coaxial line, the capacitor is adjusted with an insulated tuning tool at a frequency near the center of the band. A deep minimum in the reflected power should be obtainable. I used a Bird Wattmeter; on the 25-watt scale, using 25 watts of transmitted power, my reflected power between 144 and 148 MHz did not exceed 0.3 watt. This corresponds to a maximum VSWR of 1.25. When the plastic cover is put back in place, the tuning is hardly affected. From past experience, I know that with a more sensitive directional coupler and a little more tweaking, I could reduce the reflected power at the center of the band to practically zero. However, at present, it's close enough.

positioning mobile whip antennas

A monopole antenna is really an abstraction implying a single conductor extending out from an infinite conducting plane. Its most familiar practical implementation is a broadcast antenna tower, in which an RF voltage is applied between the base of the tower and the earth. Even here, one could regard the antenna as a very unbalanced dipole, one side being the tower and the other the earth. In mobile whips, the dipole consists of the whip

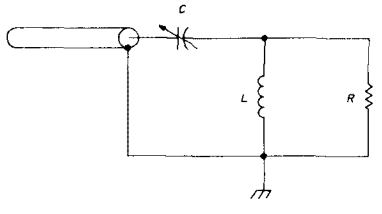


fig. 3. Schematic diagram of L-network tuner.

and the car, between which RF drive is applied.

It's a common mistake to think that the radiation pattern of a given antenna, such as a 1/4-wave, 1/2-wave, or 5/8-wave whip, is the same when mounted on a vehicle as when it's mounted above an infinite conducting plane. At 146 MHz the car's roof is a small ground plane elevated about 3/4 wavelength above the earth ground. This elevation affects the whip antenna radiation pattern.

When VHF antennas are mounted on cars, the main differences between antennas show up in the lack of symmetry of their respective radiation patterns in the horizontal plane. The most circular pattern is always obtained with the whip mounted at the center of a metal-topped roof. Contrary to popular belief, little gain differences are noted between 1/4-, 1/2-, and 5/8-wave whips when they're mounted on finite ground planes (car roofs), certainly not the 3-dB improvement realized when comparing a 1/4-wave to a 5/8-wave vertical working against an infinite-conductivity, infinite-extent ground system. (*This is related to the fact that at low take-off angles, as in mobile-to-mobile contacts, none of these antennas produces a maximum field component. It would be necessary to tilt the car and antenna system as a unit, adjusting the incline for each antenna/ground optimum take-off angle, in order to properly compare maximum developed field strengths. — Ed*)

feedpoint currents

At very asymmetrical locations, such as at one side of a trunk lid, or on one side of the hood, rather large differences between different antennas may appear. This is because of the different levels of current that emerge from the feedpoint at the bases of different antennas and spread out over the surface of the car. Of the three lengths of whips popularly used, the 1/4-wave whip spreads the most current; the current from the 5/8-wave whip is about 70 percent of the value of the 1/4-wave whip, and the current from the 1/2 wave is the lowest, being in the order of 20 percent of the value of the 1/4 wave (depending on the diameter of the whip). These differences in current spreading out over the car have little effect when the antenna is placed at a point of symmetry. However, when the whip is in a highly asymmetrical location, the patterns can be drastically affected.

radiation intensity measurements

Some years ago a group of professors at the University of Washington (including myself), who were also licensed Amateurs, made a series of measurements of the radiation intensity in the horizontal plane from 2-meter whips mounted on vehicles. The cars were mounted on a rotator, flush with the ground, and the radiation intensity was measured using a distant pick-up antenna in conjunction with a Scientific Atlanta pattern recorder. The best antenna location, as expected, was the center of the roof, where the maximum variation in intensity throughout 360 degrees of rotation was about 3 dB. The worst case was for a 1/4-wave whip mounted at the side of the trunk of a sedan, just forward of the hinge. In this case, the pattern was very irregular, showing a 17-dB hole in one direction. The most surprising result was in the case of a 1/2-wave whip mounted on the centerline of the top at the rear end, just above the tailgate window of a station wagon. The pattern for this case had a maximum-to-minimum variation of only 3 dB in 360 degrees.

There seems to be little doubt, then, that the 1/2-wave base-driven whip is the most tolerant of mounting location. It is therefore an excellent choice for a magnetically mounted antenna, especially when placed beside the trunk lid or to one side of the hood, just ahead of the windshield. Since maximum current occurs halfway up the whip, the 1/2-wave antenna can "see" over the car top better than a 1/4-wave whip at the same location.

acknowledgment

I want to thank my friend and colleague, Eric Lindahl, WN7WNL, who first noticed this antenna, bought one, and converted it for 2-meter use.

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VHF/UHF WORLD

Joe Reisert W1JR

RF connectors: part II

Last month's column introduced the subject of RF connectors, described some of the major series, and provided relevant mechanical and electrical data. This month's column will complete the subject and present additional data on the mechanical and electrical aspects of the RF connectors most often used by Amateurs.

mechanical considerations

RF connectors come in many types, shapes, and sizes, depending on whether the cable to be connected has a braided, solid, or helical/annular outer shield. The most common RF connectors are designed for flexible coaxial cable. Hardline and Heliax™ cables require special connectors which are usually bigger and more difficult to attach, frequently requiring the use of a hacksaw or tubing cutter. Crimp-type connectors, which are becoming very popular, will be discussed later.

RF connectors are manufactured from many different materials. The most common materials in the outer conductor are brass, stainless steel, aluminum, die-cast zinc, or beryllium copper. Stainless steel, one of the most durable materials, is expensive. Aluminum, usually cheapest, is the least durable.

Connector plating is very important, especially on active metals. Even stainless steel connectors are usually plated. Until the price of precious metals soared, the most common platings were silver or gold; nowadays, less expensive platings such as nickel, tin, or chrome are used.

Another important but frequently overlooked indicator of quality in an RF connector is the material of which the delicate center pin is made. The most common materials — brass, beryllium copper, and phospor bronze — are preferred because they can be readily soldered. Center pins are usually plated with a precious metal.

Depending on the connector series and type, the center pin can be captivated, connected to the center conductor of the cable, or actually fabricated from the center of the cable itself. If the center pin is captive, proper positioning inside the connector body will usually be assured. However, if the center pin is connected directly to the cable, its position on the center conductor is very critical during the time when the connection is assembled. If it's too long, it may protrude too far and damage the mating connector. If it's too short, contact may be either lacking or intermittent.

Because the center pin may move in or out with changes in the environment (such as temperature, for example), the latter can be a problem at temperature extremes or when a cable is coiled up. This is why connector manufacturers usually provide detailed assembly instructions indicating exact cutting lengths on all the sections of the cable being prepared to mate with a connector. Furthermore, coaxial cables should *always* be secured to a tower or mast so that no pressure is placed on the connector itself.

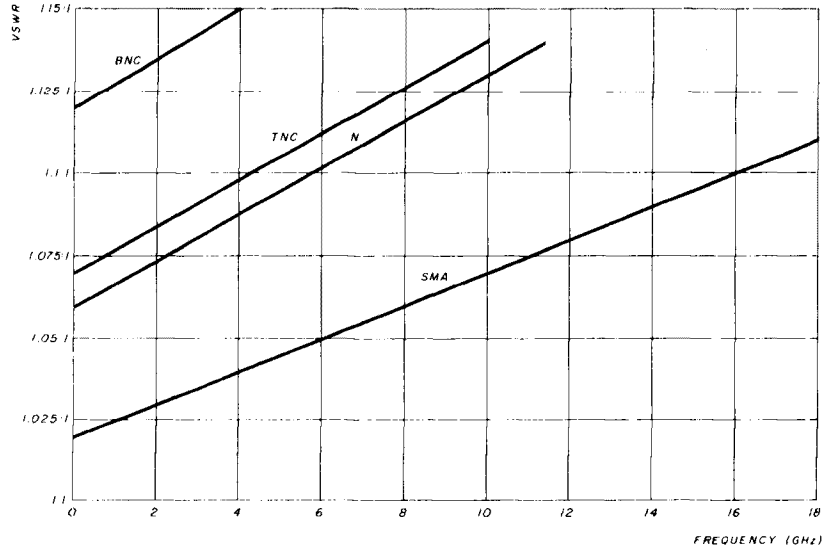


fig. 1 Typical maximum VSWR versus frequency for type BNC, TNC, N and SMA conncectors.

The quality of the RF connector's threads is also important. Although the threads are usually quite durable, those on less expensive connectors — such as the UHF type — generally don't work well after many matings. If the connector body is aluminum, the threads may seize; sometimes a lubricant is necessary.

Before mating connectors, always make sure that the connectors are properly aligned to avoid cross-threading. Sometimes it helps to first rotate a connector either one half or one full turn in the *reverse* direction before turning it in the desired direction. Finger-tightening is usually sufficient; if further tightening is required, use a wrench only as a last resort, applying only as much pressure as is absolutely necessary. Too much pressure may damage the connector permanently.

Connectors use many types of dielectrics, including air or polystyrene, teflon, Rexolite™, glass-teflon, and mica-filled bakelite. Because these materials often require a precise fit, they must be properly handled when connectors are assembled. Any heat (such as that applied by a soldering iron during assembly) or external side-pressure can permanently distort the dielectric.

It should go without saying that all connectors must be properly assembled, especially if they're to be used near their frequency limits. Designers have spent countless hours ensuring that connectors will meet their specifications; if improperly assembled, the specified impedance or insertion loss may be exceeded.

One frequently overlooked mechanical parameter of RF connectors is the quality of the environmental seal. Is the connector weatherproofed? Does it have a gasket or washer made of synthetic rubber, silicon rubber, or neoprene? If it doesn't, it should never be used outdoors or in indoor applications where moisture is present.

electrical performance

The most important electrical properties of an RF connector are frequency of operation, impedance, power handling capability, and insertion loss. Reference 1 details the typical maximum frequencies for the major series of connectors. Above the maximum recommended frequencies, the impedance of an RF connector may wander and "suckouts" attributable to resonances may occur. This can severely limit performance, perhaps without your knowledge.

Connectors may be damaged if subjected to high RF levels, especially if there's a VSWR mismatch. VSWR performance is usually tightly controlled on RF connectors; for UG-type connectors, it's generally spelled out in the various military standards. Remember that this maximum VSWR applies *only* if the connector is properly assembled to a low-VSWR cable. If the cable has a poor VSWR, or if the connector is improperly assembled, all bets are off.

VSWR usually increases with frequency at a fairly predictable rate below the cutoff frequency of the connector. It can be defined mathematically by the following equation:

$$VSWR = K_1 + K_2 f \qquad (1)$$

where K_1 and K_2 are VSWR constants (that depend on the connector series) and f is the operating frequency in GHz. The lower the constant, the lower the VSWR.

The typical VSWR constants for the RF connector series most popular with Amateurs are shown in **table 1**. For instance, using **eqn. 1**, a properly assembled N connector should have a VSWR below 1.088 at 4 GHz and 1.13 at 10 GHz. **Figure 1** shows some typical connector VSWR variations with frequency.

Insertion loss is another important RF connector parameter. The maximum insertion loss of a connector can be estimated using **eqn. 2**:

$$I.L. = K\sqrt{f} \qquad (2)$$

where $I.L.$ is insertion loss in dB, K is the insertion loss constant from **table 1** and f is frequency in GHz. For instance, using the constants in **table 1**, the maximum insertion loss of a properly assembled N connector at 4 and 10 GHz will be 0.1 and 0.158 dB, respectively. **Figure 2** has been prepared to show some typical losses of RF connectors at different frequencies.

In last month's column I alluded to the fact that some Amateurs throw connector specifications around quite loosely. For example, I'll bet most of you have heard somebody say that connectors should be avoided at all cost since a typical RF connector has up to 0.5-dB insertion loss. A 0.5-dB insertion loss is equal to a power dissipation of 11 percent. Said another way, if 1000 watts of RF power were

Table I. Parameters needed to properly choose an RF connector in the main series used by Amateurs. Typical constants for determining VSWR and insertion loss versus frequency are also shown where known. Consult text for explanation on use of the constant.

Connector series	Peak RF Voltage at 100 MHz	Peak power in watts	Impedance in ohms	Max freq in GHz	VSWR constants	Max insertion loss constant	RF leakage in dB**
BNC	1000	500	50*	4	1.12 + 0.007f	0.1	−60 + f
N	1000	1000	50*	11	1.06 + 0.007f	0.05	−85 + f
SMA	1000	500	50	18	1.02 + 0.005f	0.03	100 + f
TNC	1000	1000	50	11	1.07 + 0.007f	0.05	−90 + f
UHF	500	500	50-70	0.3	NA	NA	NA

*75-ohm versions are in use, especially in the CATV and computer industries.
**f in GHz (see text for further explanation).

to pass through a connector with a 0.5-dB insertion loss, 110 watts would be dissipated. *Goodbye connector!* Even a 0.1-dB loss is about 2.25 percent, or, under the same conditions, a 22.5-watt dissipation. This is still enough loss to destroy most RF connectors when sufficient power is applied.

Another important RF connector parameter is power handling capability. It's painfully obvious from the material just presented that insertion loss, to a great extent, is the determining parameter when specifying power. Indirectly linked to power is the breakdown voltage of the connector.

Power handling is inversely related to frequency. Hence the power rating of an RF connector decreases with increasing frequency. **Figure 3** has been prepared to show typical manufacturers' ratings of RF power versus frequency for some of the most popular

Most Amateurs running high power on 70 cm and above have probably exceeded the ratings on N connectors at one time or another even with the old FCC power limits. Furthermore, the power limits assume a 1:1 VSWR. What do you do if your VSWR isn't 1:1? You must derate the power handling capabilities of the connector accordingly.

Table 2 shows the recommended percentage of RF power derating for an RF connector with a VSWR greater than 1:1. For instance, if your VSWR is 2:1, the connector power rating shown in **fig. 3** should be decreased to 88 percent. Therefore, the maximum rating of an N connector at 1:1 VSWR operating at 432 MHz is approximately 450 watts. With a 2:1 VSWR at the same frequency, the maximum power rating should be dropped to approximately 400 watts!

The voltage breakdown on RF connectors isn't always clear. Manufacturers usually do a high potential breakdown, but this is at DC. The actual RF breakdown, typically done at 5 MHz, is much lower. Some typical limits are shown in **table 1**. *Note that in this* regard there's normally *no difference* between BNC, N, TNC or SMA-type connectors; therefore, size isn't the only determining parameter of the RF breakdown voltage.

Table 2. This table lists the recommended deratings of a connector's RF power handling if the VSWR is higher than 1.1:1.

VSWR	Derating factor (in percent)
1.5:1	94
2.0:1	88
2.5:1	83
3.0:1	78
3.5:1	73
4.0:1	70
4.5:1	67
5.0:1	65

The VSWR indirectly enters the picture because the voltage across the connector may rise if the VSWR isn't 1:1. As noted in reference 1, some series of RF connectors are highly recommended if breakdown voltage is a criterion.

RF leakage is sometimes a criterion, especially if the connector is operated in a strong field (such as the output of a down-converter feeding a 28 MHz I-F.). Typical values of RF leakage are shown in **table 1**, where *f* is frequency in GHz. Hence a BNC connector at 4 GHz will have a leakage of only 56 dB below the internal or external signal level, while an SMA connector at the same frequency will have a leakage of typically 96 dB!

adapters

So far we've been discussing only RF connectors. RF adapters are also very common. Generally speaking, they're about two to three times poorer than a single connector. Therefore, to estimate the VSWR or insertion loss of a coaxial adapter, double or triple the value shown for a single connector.

crimp connectors

Although most Amateurs use standard RF connectors that can be assembled with simple hand tools, crimp-type connectors have been available for many years. These connectors, however, weren't always reliable, were often more costly than comparable non-crimp types, and usually required expensive crimping tools.

In recent years this situation has changed. Crimp-type connectors have become less expensive and more reliable; any necessary crimping tools have also decreased in price. In fact, Gilbert Engineering* has developed type-N crimp connectors for RG 8/U, RG 213/U, RG 214/U, and Belden 9913 that are less than two-thirds the price of comparable non-crimp-type connectors. They all use the same crimping tool, which costs about $45.00.

Once you've assembled about 30 crimp-type connectors, the crimp tool

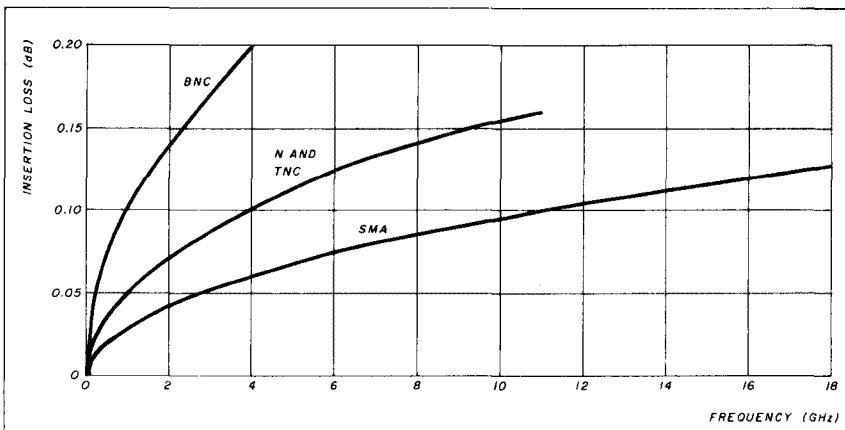


fig. 2. Typical maximum losses versus frequency for some popular rf connectors used by amateurs.

*Gilbert Engineering Co., P.O. Box 23189, Phoenix, Arizona 85063-3189

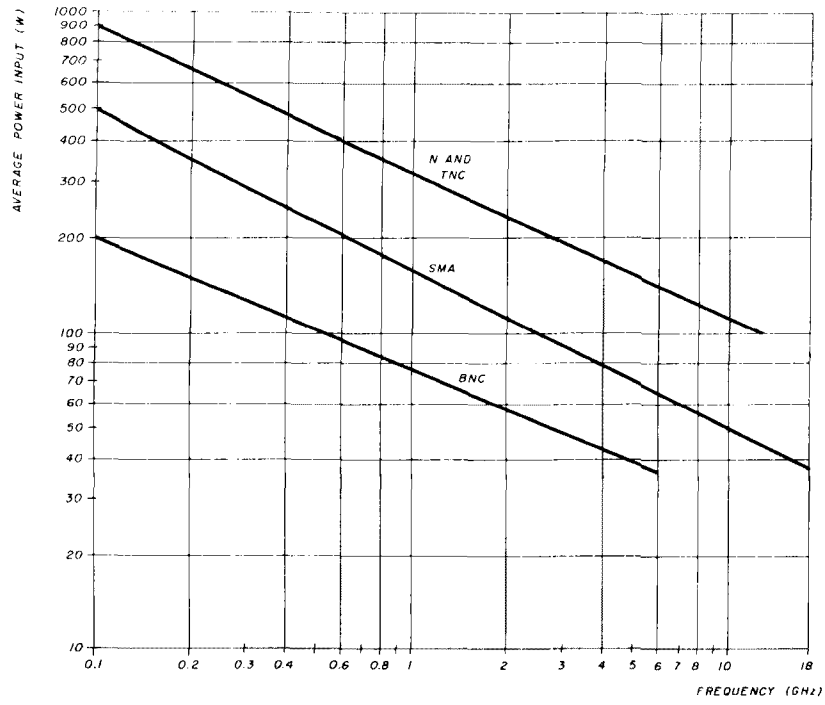


fig. 3. Manufacturers rf power ratings for some of the most popular rf connectors. Note how they drop rather rapidly at the higher frequencies.

has paid for itself and you'll save money on each connector thereafter — not to mention the considerable savings in assembly time. I've used all Gilbert Engineering types mentioned and can attest to their performance and ease of assembly. No more tight trimming or filing down the center of the coax!

The only problem with most crimp-type RF connectors is that they're not waterproofed. However, if you slip a small (2- to 3-inch) piece of heat-shrinkable tubing over the rear portion of each connector, you'll have a very weatherproof connector that probably has a better seal than the typical non-crimped connector with its built-in environmental seal. These connectors should be especially useful in alleviating the drudgery of assembling a large antenna array!

availability and cost

Good new RF connectors aren't cheap. The price of most common connectors runs from $2 to $5 each in small quantities. The smaller the connector and the higher the frequency of the operation, the more costly they're likely to be. Likewise, connectors with silver or gold plating or teflon dielectric are more expensive. If you're going to use a connector outdoors, make sure it has a weatherproof seal.

Connectors for hardline and Heliax™ are much more expensive. This is easy to understand because they're bigger, more complex, and more difficult to manufacture. However, once they're installed on a line, they'll give many years of service and will pay for themselves in improved performance.

Don't skimp on connectors, especially if you use an expensive transmission line. Low-cost connectors could break down or allow moisture to form inside the connector and thereby ruin an expensive transmission line. Do shop for bargains, especially at flea markets, but don't be penny wise and pound foolish.

tips

Always be sure that a neoprene washer or equivalent seal is used on all connectors mounted outdoors. Sealing connectors from the weather is tricky business. Some Amateurs recommend special sealing techniques on the connectors, while others say not to worry about any of the above. I suggest taking a position somewhere in between; if you try to seal off a connector completely (using tape, acrylic strays, or rubber/silicon-type compounds) and *any* moisture enters (and believe me, it will), you're in real trouble. I've tried placing silicon grease inside RF connectors only to have it degrade with time and then take on moisture; this can result in an explosion, however, if the VSWR gets too high!

If you do decide to seal a connector, Coax-Seal™ is highly recommended. It's reasonably priced and does a good job of filling most of the places where moisture can enter. It's also reworkable if the connector has to be opened. But unsealed RF connectors aren't always as much of a problem as you might think. If water enters a UG21/U type-N connector, it will frequently evaporate in a day or so.

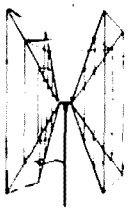
It's best to position connectors so that rain is less likely to enter and then take reasonable precautions. For example, moisture will often creep along the threads of a connector via capillary action. You can prevent this by applying a light coating of silicon grease to the exposed threads of one of the connectors just before final connection to the antenna. Once the connector is screwed together, the protective coating will prevent moisture from entering.

As mentioned earlier, RF connectors should never be subjected to pushing or pulling. Mount cables securely with tape, Tiewraps™, or Kellum clamps ("torture fingers") so that no external pressure is exerted on the connector. This is particularly true with RG 17/U coax, which has a different expansion coefficient between the center conductor and the shield. Tie RG 17/U down carefully, *without bends*, to prevent pin pullout in cold weather.[2]

It's important to pay close attention to the assembly of an RF connector so that the center pin is properly located. (This is especially important with the

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"captivated" type of connectors.) One test I perform involves placing a fingertip on the center pin of the assembled connector. If the finger lightly touches the pin, then the pin is probably assembled properly. If the pin tries to puncture your fingertip, it's probably protruding too far. Conversely, if you can't feel the pin, then it's too far inside the connector body and probably won't have sufficient overlap for a reliable connection.

For maximum reliability, test all connectors for VSWR after assembly. If possible, also check insertion loss. It's a lot easier to test and repair an RF connector in the shack than on an antenna mounted 100 feet in the air!

summary

RF connectors are constantly being improved, and Amateurs who follow these developments will reap the benefits. Those who take the time to properly assemble connectors — following the manufacturer's instructions — will have the least amount of downtime and the most pleasure in operation.

With the data provided in last month's and this month's columns, you should have all the information necessary to choose the right RF connector for your application. Remember to use a connector type that will have low loss and be able to handle the RF power required at the frequency of operation.

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acknowledgments

I'd particularly like to thank Charlie Button of Gilbert Engineering for his helpful suggestions with the use of crimp-type connectors.

references

1. Joe Reisert, W1JR, "VHF-UHF World: RF Connectors, Part I," *ham radio*, September, 1986, page 77.
2. Joe Reisert, W1JR, "VHF-UHF World: Transmission Lines," *ham radio*, October, 1985, page 83.

Important VHF/UHF events:

October 4-5:	*Mid-Atlantic States VHF Conference, Warminster, Pennsylvania (contact WA2OMY)*
October 4-5:	*International Region 1 UHF/SHF Contest*
October 7:	*EME perigee*
October 9:	*Predicted peak of the Draconids Meteor Shower at 0900 UTC*
October 10-11:	*First ARRL 10 GHz Cumulative Contest, second weekend*
October 20:	*Predicted peak of the Orionids Meteor Shower at 0900 UTC*
October 25-26:	*ARRL EME contest, first weekend*
November 2:	*Predicted peak of the Taurids Meteor Shower at 1520 UTC*
November 3:	*Predicted peak of the Cassiopeids Meteor Shower at 1515 UTC*
November 4:	*EME perigee*
November 17:	*Predicted peak of the Leonids Meteor Shower at 0850 UTC*
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transistor substitution: part II

Last month we discussed the use of crossover guides, derating specs for mobile and other over-temperature applications, and certain other matters. In this second and final installment we'll look at frequency response and mechanical problems with replacement transistors.

frequency response

Suppose you've installed a "universal" small-signal replacement transistor and it doesn't amplify. According to the DC voltages, however, it draws normal current and doesn't heat excessively — yet still fails to operate. Chances are good that the transistor selected has a bandwidth that's too narrow, and so the amplification is insufficient.

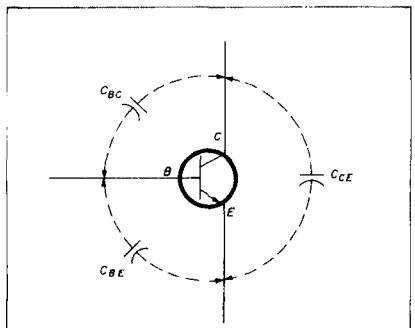


fig. 1. Transistor junction capacitances affect frequency response.

As you know, there's no one way of measuring the frequency response of transistors. Worse yet, the manufacturers can't seem to agree on the correct method of rating the frequency response. In fact, I've seen examples of three different methods all used in the same crossover guide!

Probably that's the reason you can

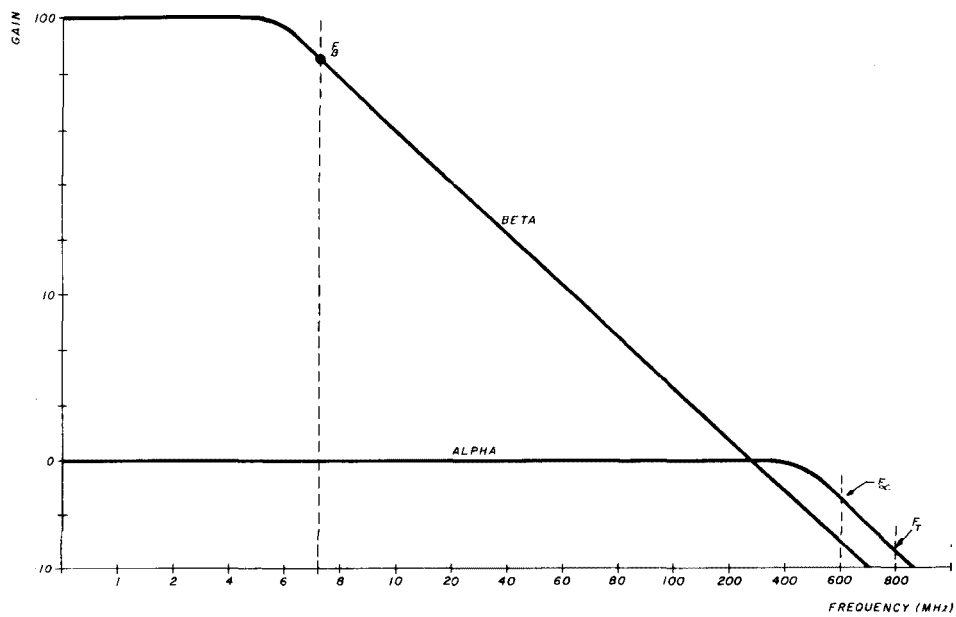


fig. 2. Typical Alpha and Beta current gains versus frequency.

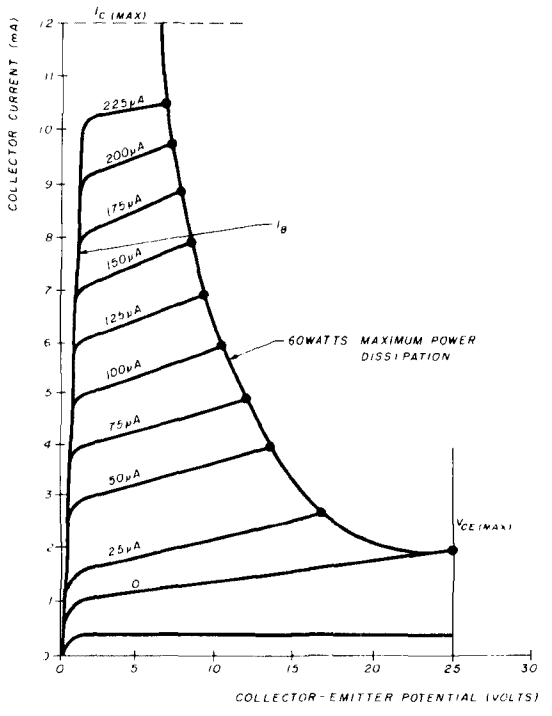


fig. 3. Maximum voltage, current and power dissipation ratings are related and must not be exceeded.

install a transistor rated "50 MHz" and yet find it won't properly amplify a signal in a 10.7-MHz IF amplifier in a 2-meter FM radio receiver. In rating the transistor at 50 MHz, the manufacturer wasn't lying — but he may have inadvertently used a rating system that doesn't fit your circuit.

As shown in **fig. 1**, one factor in rating transistors is the capacitances between the junctions. Another factor is the thickness and geometry of the base region and the time necessary for the majority of the electrical carriers to cross it. If the capacitances of the replacement are too far out of tolerance, then the gain will be reduced at high frequencies and any LC tank circuits in the transistor amplifier will be mistuned.

There's also the Miller effect, which is the effective capacitance produced by internal feedback from the output signal back to the input "amplifying" the actual capacitance. Because of the Miller effect, a small difference in internal capacitance can create a larger change in effective capacitance.

alpha and beta gain

Transistor alpha gain, the ratio of the collector current to the emitter current, can never exceed 1. Alpha gain is often measured in a common-base circuit. Beta gain, the ratio of collector current to base current, is usually measured in the common-emitter configuration. As illustrated in **fig. 2**, the frequency cutoff point is greatly different for these two kinds of ratings. A manufacturer who rates transistors by the common-base method might correctly list them as having a far wider response than would be possible if the common-emitter method were used. Of course, the common-base rating can provide a no-gain 10.7-MHz IF stage!

gain-bandwidth product

Another way of rating frequency response is the gain-bandwidth

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MRF422*		150W	38.00	82.00
MRF426,/A*		25W	18.00	42.00
MRF433		12.5W	12.00	30.00
MRF449,/A	Q	30W	12.50	30.00
MRF450,/A	Q	50W	14.00	31.00
MRF453,/A	Q	60W	15.00	35.00
MRF454,/A	Q	80W	15.00	34.00
MRF455,/A	Q	60W	12.00	28.00
MRF458		80W	20.00	46.00
MRF475		12W	3.00	9.00
MRF476		3W	2.75	8.00
MRF477		40W	11.00	25.00
MRF479		15W	10.00	23.00
MRF485*		15W	6.00	15.00
MRF492	Q	90W	16.75	37.50
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MRF222	25W	136-174	14.00	—
MRF224	40W	136-174	13.50	32.00
MRF237	4W	136-174	3.00	—
MRF238	30W	136-174	13.00	30.00
MRF239	30W	136-174	15.00	35.00
MRF240	40W	136-174	18.00	41.00
MRF245	80W	136-174	28.00	65.00
MRF247	75W	136-174	27.00	63.00
MRF260	5W	136-174	7.00	—
MRF261	10W	136-174	9.00	—
MRF262	15W	136-174	9.00	—
MRF264	30W	136-174	13.00	—
MRF607	1.75W	136-174	3.00	—
MRF641	15W	407-512	22.00	49.00
MRF644	25W	407-512	24.00	54.00
MRF646	40W	407-512	26.50	59.00
MRF648	60W	407-512	33.00	69.00
SD1441	150W	136-174	74.50	170.00
SD1477	100W	136-174	32.50	78.00
2N3866*	1W	30-200	1.25	—
2N4427	1W	136-174	1.25	—
2N5591	25W	136-174	13.50	34.00
2N6080	4W	136-174	7.75	—
2N6081	15W	136-174	9.00	—
2N6082	25W	136-174	10.50	—
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2N6084	40W	136-174	13.00	31.00

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product method, defined as the frequency at which the common-emitter gain drops to unity. For example, let's look at a transistor with a low-frequency (1000 Hz) beta rating of 50 and a gain-bandwidth product of 50 MHz. The gain-bandwidth product equals beta times the common-emitter cutoff frequency, so the common-emitter cutoff frequency is found to be only 1 MHz. That's why you can act on some of the short-form specifications and still obtain a dud that won't amplify.

analyzing maximum ratings

Care must be used in analyzing the manufacturer's maximum voltage and current ratings. Just because a transistor is listed for certain maximum collector voltage and current doesn't mean it can always operate safely at those levels. **Figure 3** shows that the transistor can safely stand either high current or high voltage, but not both at the same time. Maximum collector dissipation (watts), the *product* of both

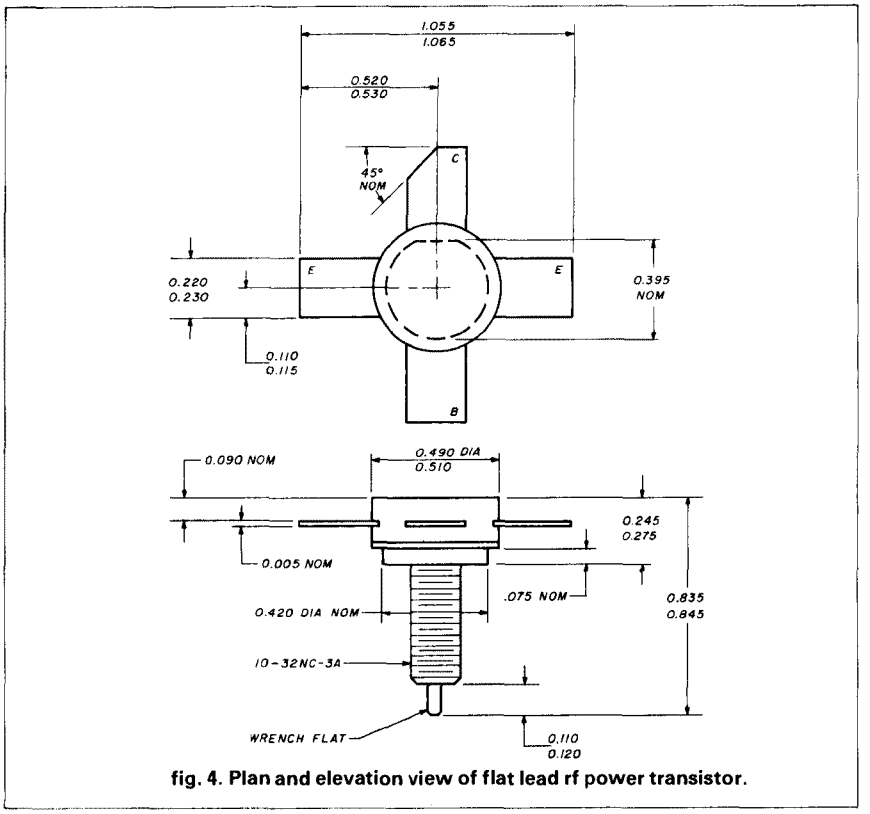


fig. 4. Plan and elevation view of flat lead rf power transistor.

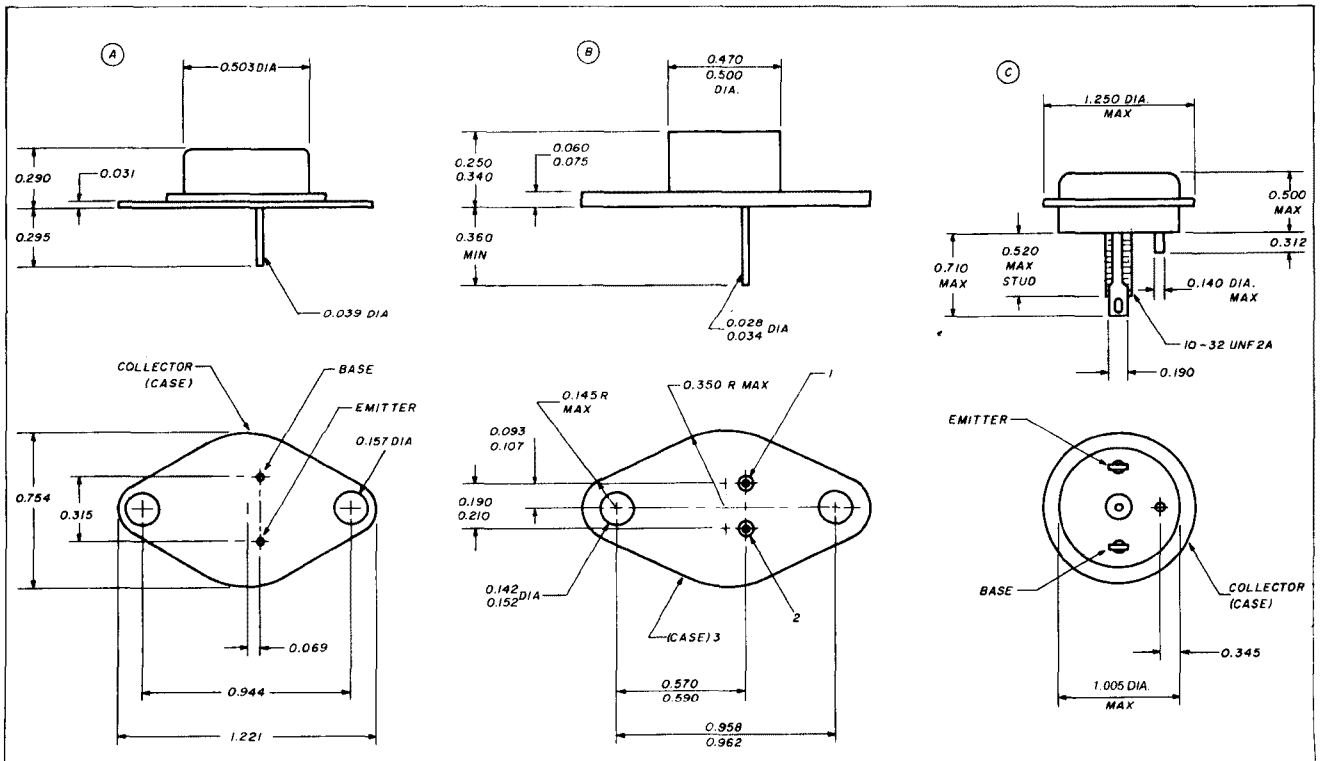


fig. 5. Power transistor configurations (A) TO-3, diamond shaped package; (B) TO-66; (C) TO-36.

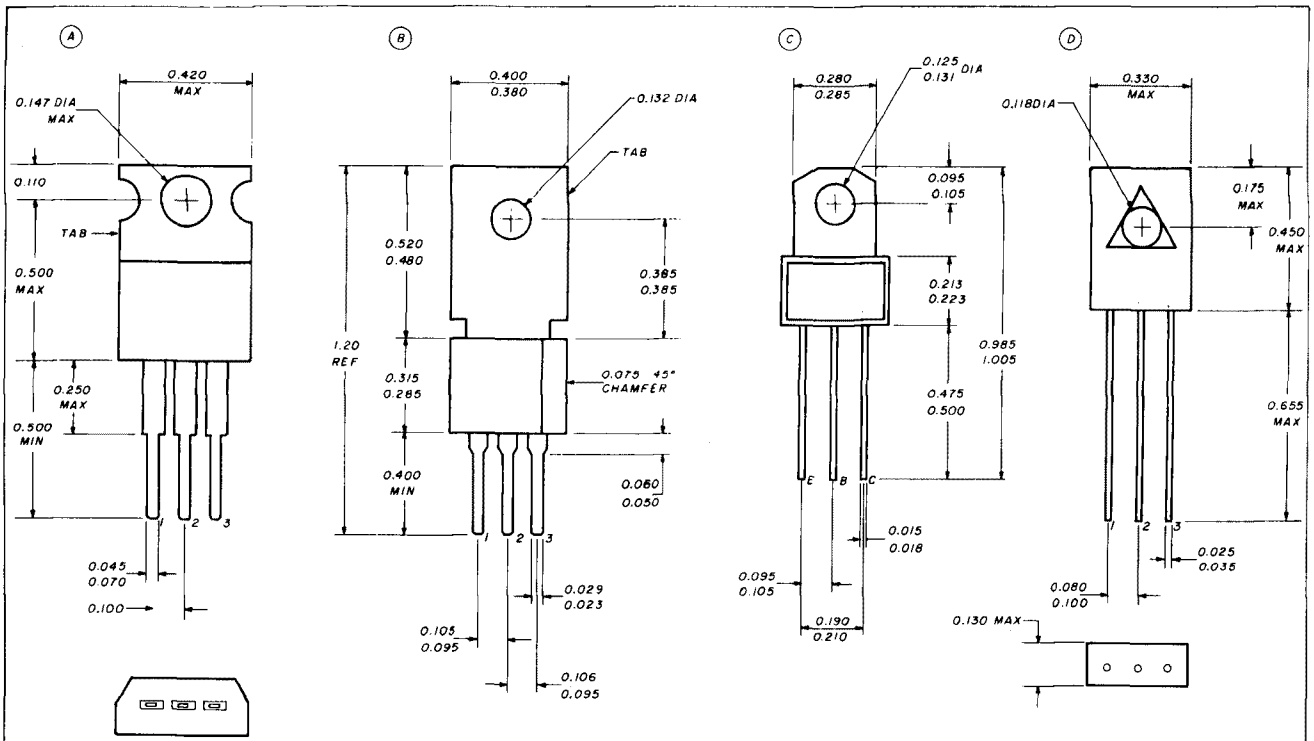


fig. 6. Popular plastic transistor configurations: (A) TO-200; (B) TO-202; (C) another tab-mounted package; (D) mounting hole through body of transistor.

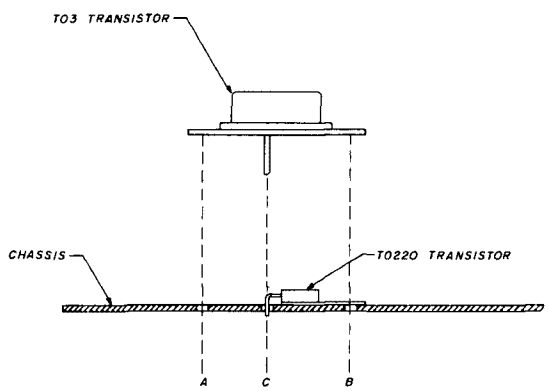


fig. 7. Modifying a TO-220 transistor for TO-3 replacement.

voltage and current, must not be exceeded.

mechanical problems

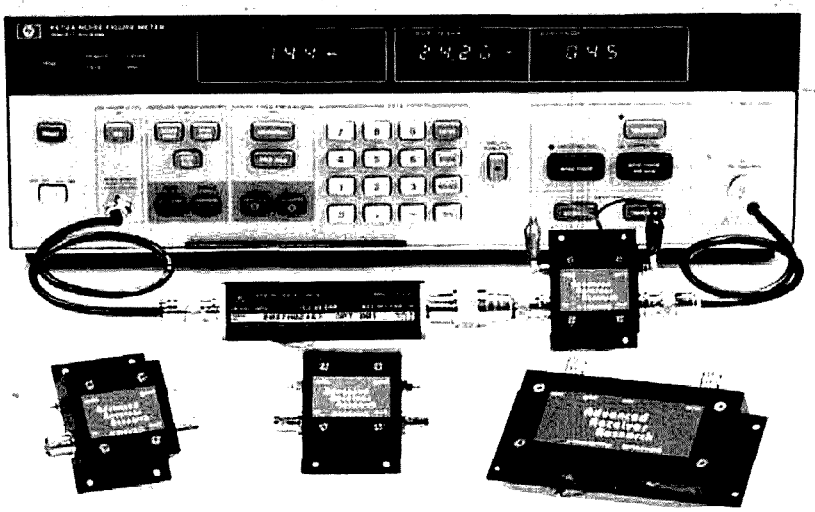
Problems of physical size, connecting leads, and methods of mounting some substitute transistors are equally as vexing as those of finding a suitable electrical characteristic.

Our first transistor package is the RF power transistor shown in **fig. 4.** This device uses thin, flat "low inductance" leads. Several sizes are available, and size doesn't necessarily indicate relative power dissipation rating (although it usually does). There was one kit-form 2-meter FM mobile power amplifier that came with either of two different types of power transistor. The hole in the printed wiring board was cut for the larger type and rubber "O-rings" were placed around the smaller to make them fit.

Figure 5 shows several types of power transistor package. The TO-3 transistor in **fig. 5A** is the so-called "standard" power transistor in a diamond-shaped package. A smaller diamond-shaped package is the TO-66 shown in **fig. 5B**. There's also a Japanese "similar-to-TO66" package that looks, at first glance, like the TO-66 but has slightly different pin spacings. Finally, the big horse shown in **fig. 5C** is the TO-36. This high-power transistor is used extensively in automotive audio applications, mobile solid-state HV multivibrator DC power supplies, and in industrial electronics applications.

Figure 6 shows several popular plastic power transistor packages. Some of these are listed as replacements for TO-3 or TO-66 diamond-

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SP144VD	144-148	<1.6	15	0	DGFET	$59.95
SP144VDA	144-148	<1.1	15	0	DGFET	$67.95
SP144VDG	144-148	<0.55	24	+12	GaAsFET	$109.95
SP220VD	220-225	<1.9	15	0	DGFET	$59.95
SP220VDA	220-225	<1.3	15	0	DGFET	$67.95
SP220VDG	220-225	<0.55	20	+12	GaAsFET	$109.95
SP432VD	420-450	<1.9	15	−20	Bipolar	$62.95
SP432VDA	420-450	<1.2	17	−20	Bipolar	$79.95
SP432VDG	420-450	<0.55	16	+12	GaAsFET	$109.95

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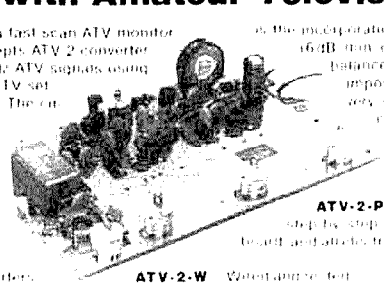
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shaped power transistors. The package in **fig. 6A** is the TO-220 (once also called "P-66") and is common in small, low- to medium-powered audio applications and most car radios. Two additional tab-mounted plastic power transistors are shown in **figs. 6B** and **6C**. Finally, the device shown in **fig. 6D** represents a class of Motorola power transistors that are not tab-mounted, but instead have a mounting hole through the body of the transistor.

Figure 7 shows how a plastic, tab-mounted TO-220 power transistor can replace a TO-3 power transistor. The center terminal (collector) is cut off the TO-220; it won't be needed because the tab mount is also connected to the collector. The base and emitter leads are bent down and the mounting screw is passed through the tab hole into the original mounting hole for the TO-3 transistor.

conclusion

Ideally, when a replacement transistor is needed we'll have an original from the manufacturer of our rigs on hand or easily available. But sometimes we're either unable to obtain such a transistor or the cost is prohibitive. In such cases we can usually make an educated guess at a proper replacement type.

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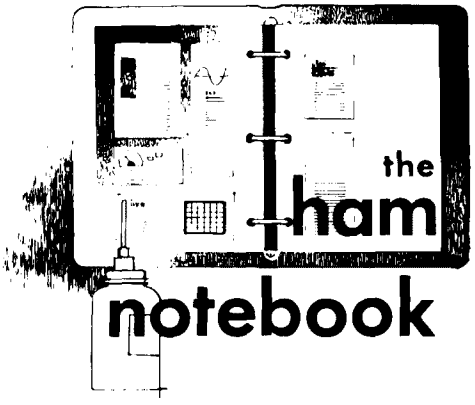
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matching linears to transceivers on 10-15 meters

Grounded grid linear amplifiers don't always present a perfect 50-ohm resistive input for the transceiver. And long leads from the transceiver sometimes present a reactive load on the 10- and 15-meter bands. What happens? The transceiver goes into self-oscillation when the amplifier is driven.

To solve this problem I first tried a transmatch, but it didn't work because of the difficulty of obtaining short leads. An "L" network didn't work, either.

But the "T" match (**fig. 1**), with its short leads, worked the first time. The transceiver loaded up to full power, driving the linear with no ensuing oscillations. I was delighted with the results.

The matcher was built in a small LMB Flange Lock Box (No. EL532), measuring 5¼ x 3 x 2½ inches (**fig. 2**). With the capacitor bolted to the front panel and the two input and output BNC coax fittings on the back, leads were kept short. The capacitor should tune 10 meters with a 50 pF setting and adjustment coil. For 15 meters I just increased the setting to about 70 pF. The SWR dipped 1:1 without any switching or shorting out turns for 10 meters. I could cover both bands.

Calculations would show about 50 pF for 15 meters and 45 pF for 10 meters, adjusting the coil for that value. However, I don't find this necessary.

Ed Marriner, W6XM

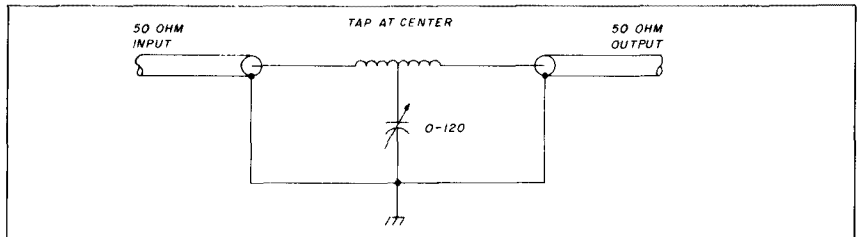


fig. 1A. "T" match circuit.

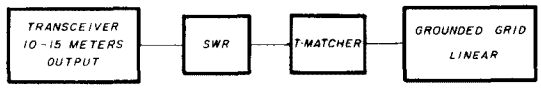


fig. 1B. "T" match block diagram.

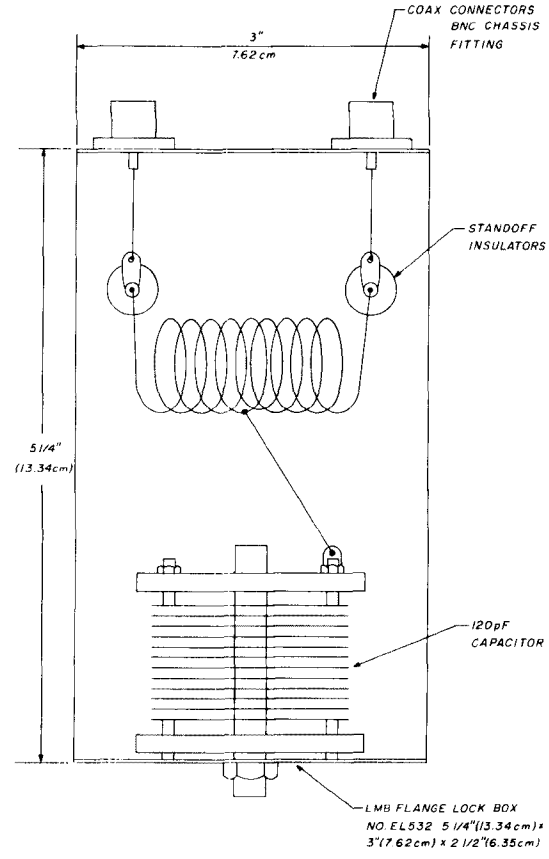


fig. 2. Details of "T" match construction.

coupling to a GG linear

Some of the grounded linear amplifiers, especially home brew jobs, capacity couple directly into the filament RFC

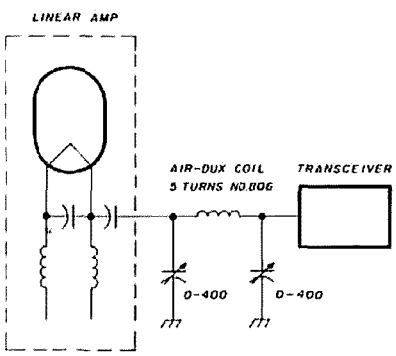


fig. 3. Network matches Swan 350B to a 3-500Z input.

through a coupling capacitor. Generally the coax link is rather long; the driver transceiver has to work a lot harder than necessary to make up tor the loss in mismatch. A 3-500Z, for example, has an input of about 150 ohms.

A simple matching device to match the output of the transceiver to 150 ohms is a pi-network. Many handbooks show a pi-network using fixed capacitors. I found being able to vary the capacitors made it easier to match. **Fig. 3.** shows a network I used for matching a Swan 350B 50-ohm output to a 3-500Z 150 input.

The unit was constructed in a small box and inserted in the coax line to the amplifier. The capacitors were small 0-400 pF BDC type, and did not arc over. The coil can be either five turns of Airdux #806 or five turns of No. 14 AWG wire wound six turns per inch on a 3/4-inch coil form. It takes very little inductance for 15 meters.

First I loaded the Swan 350B into a 50-ohm load. I then fed a signal into the amplifier and adjusted both capacitors for maximum output. Next, the loading capacitor on the Swan 350B was adjusted slightly for maximum output again on the linear. I found it took a lot less drive from the 350B to drive the linear over the old method, without a matching device. It's worth the effort to build one for each band if necessary. The received signal has to come through this network, but it does not seem to cause any deterioration in signal sensitivity.

Ed Marriner, W6XM

line regulator

It's often desirable to adjust the 120-volt AC line voltage if you're using a computer or other equipment that won't tolerate high or low line voltage.

Although line voltage could be varied with a large variac, there is a way to do this inexpensively — through a system called "line bucking" — using a small variac and a filament transformer (**fig. 4**).

In line bucking, a 120-volt line runs through the low voltage secondary of a filament transformer. This winding must carry the current of whatever you're drawing with your equipment. I used a 10-ampere, 5-volt filament transformer.

By varying the voltage on the primary side, you can add or subtract on the secondary. I used a surplus General Radio 200-B 2-ampere variac to perform this operation. By varying the arm on the variac I can adjust the line voltage by several volts plus or minus. I bring my 127-volt line down to 120 volts very nicely and don't have to worry about burning out my equipment.

The project is well worth the effort required to locate parts. I found mine — for less than $10 — at a radio swap meet and surplus store.

Ed Marriner, W6XM

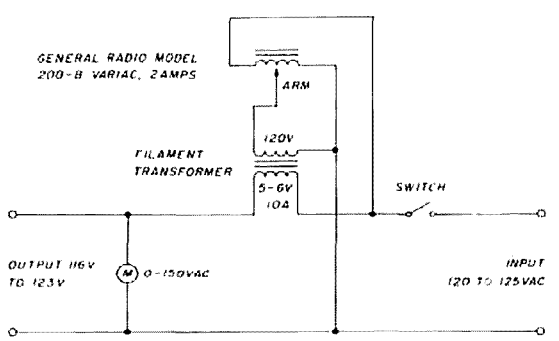


fig. 4. Combination of small variac and filament transformer provide fine line voltage adjustment.

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an easy way to measure capacitance

Like many hams, I have a can of capacitors whose values are either unmarked or marked in a code I don't understand. I use a simple method to measure their values.

The equipment required is modest: all you need is a signal generator, a frequency counter (if you don't have one, use the frequencies indicated on the signal generator) and an indicating meter together with a few resistors, capacitors, and diodes. The signal generator should cover the audio and low radio frequencies. The indicating meter can be a VOM.

Figure 5 shows the basic circuit. A signal from the generator is applied to a known resistance, *R*, connected in series with the unknown capacitor, *C*. The frequency of the signal is adjusted so that the RMS voltage across the resistance and the capacitance are equal as measured by two voltmeters, *Vac*. Then,

$$X_C = \frac{1}{2\pi fc} = R \qquad (1)$$

where:

X_C = reactance of the capacitor, ohms
f = frequency of the applied signal, Hz
C = capacitance, Farads
R = resistance, ohms

After some calculations I chose *R* to be 1000 ohms. Rearranging **eqn. 1**, substituting 1000 for the value of *R*, and changing a constant so that *C* is now in μF, we have:

$$C = \frac{159.15}{f} \qquad (2)$$

Figure 6 illustrates the actual circuit I used. Instead of measuring the AC voltage across the resistance and the capacitance, the AC is rectified by each of the two diodes and used to charge the two 4-μF capacitors. The resulting DC then is measured by a voltmeter that's connected as indicated.

Accuracy of the results will depend on the actual value of the known resistance, *R*, and your ability to determine the frequency of the applied signal. Use a high-tolerance resistor or measure it and substitute the actual value into **eqn. 1**. The values of the other resistors and capacitors aren't critical. The diodes can be any signal diodes.

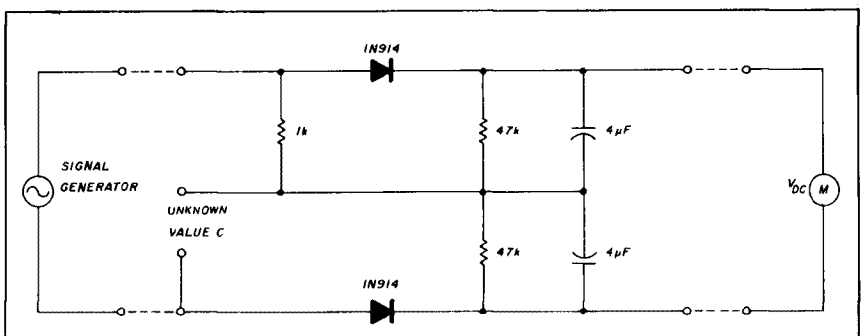


fig. 6. Simple circuit uses just one DC voltmeter.

operation

The unknown capacitor is connected and a signal from the generator is applied to the circuit. The frequency of the signal is adjusted so that the output as indicated by the DC voltmeter is zero. The frequency of the applied signal is noted (or measured if you have a frequency counter) and the capacitance is calculated from **eqn. 2**.

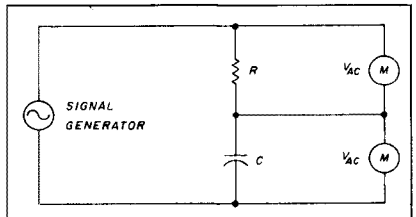


fig. 5. Equal voltage drops indicate equal resistance/reactance values.

For the DC voltmeter, I used a zero-center meter I happened to have. If you don't have such a meter, use an ordinary DC voltmeter and connect the polarity of the meter as required. Remember that when you measure a capacitor, the DC output voltage will be of one polarity when the frequency of the signal generator is too low and of the opposite polarity when the frequency is too high. You could add a polarity-reversing switch for the meter or use the bridge arrangement shown in **fig. 7**, although this scheme would broaden the frequency range for the zero-voltage indication somewhat because of the voltage drop across the bridge diodes.

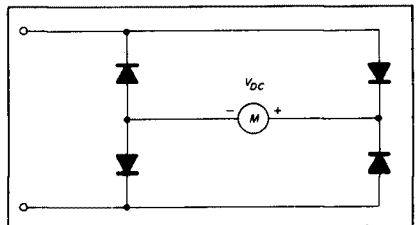


fig. 7. Bridge arrangement corrects for polarity change.

Table 1. Measurements of known capacitors.

Nominal capacitance, μF	Zero-voltage output frequency, Hz	Calculated capacitance, μF
0.05	3305	0.048154
0.02	8509	0.018704
0.01	15,761	0.010098
0.005	32,404	0.004911
0.002	82,875	0.001920
0.001	168,915	0.000942
0.000510	317,853	0.000501
0.000470	358,631	0.000444
0.000330	498,930	0.000319

performance

I built the circuit shown in **fig. 6** and checked it using some 5-percent tolerance capacitors. **Table 1** shows the nominal value of those capacitors, the frequencies at which I obtained zero-output voltage, and the calculated values of the capacitors using **eqn. 2**. The calculated values of capacitance are in good agreement with the nominal values. I didn't have any 5-percent capacitors larger than 0.05 μF, but the system should work well with larger capacitors. From **eqn. 2**, if your signal generator goes down to 100 Hz, you can measure capacitors as large as 1.59 μF. If a smaller value of *R* is used, even larger capacitors can be measured.

For my tests I used a Hewlett-Packard model 200-CD signal generator and a Heathkit model 2410 frequency counter.

A word of caution: this method measures the impedance of an unknown. For a capacitance, with the exception of electrolytics, the impedance is nearly the same as the reactance. However, this is not true for an inductance. You can measure the impedance of an inductance with this method — but unless you know the AC resistance of the inductance, you can't determine the L value of the inductor. The AC resistance is different from the DC resistance, so a measurement made with an ohmmeter won't solve that problem.

H.H. Hunter, W8TYX

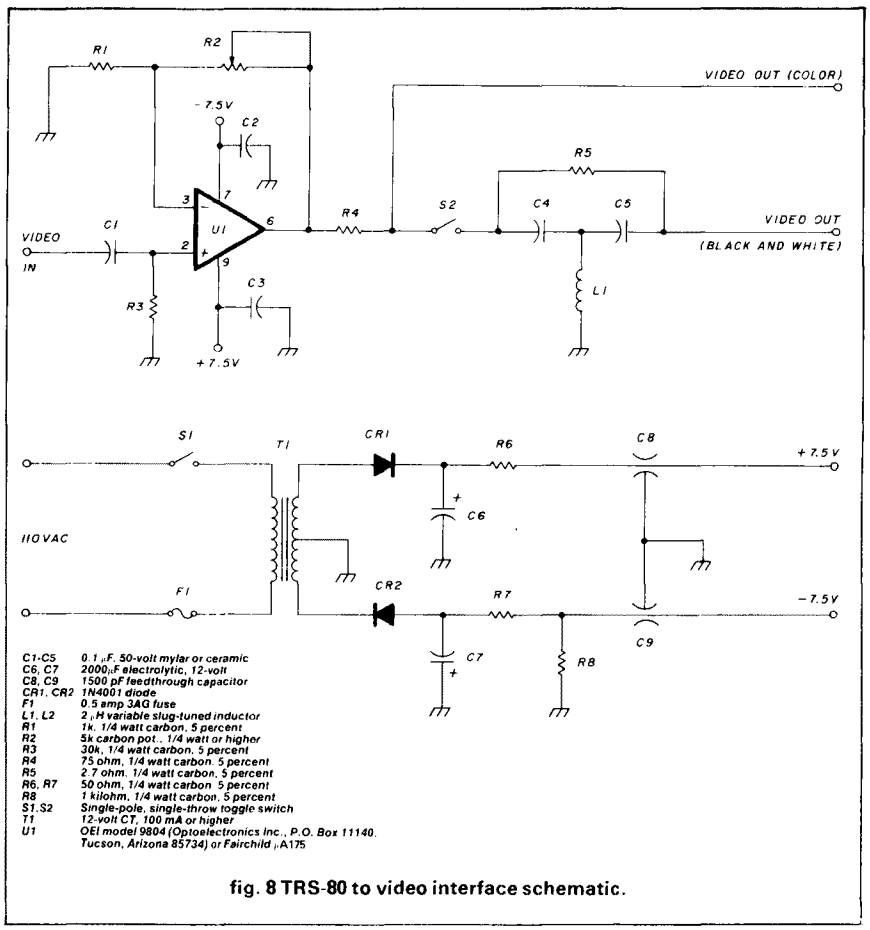


fig. 8 TRS-80 to video interface schematic.

TRS-80C to video monitor interface

This article describes a video amplifier-filter used in conjunction with a Radio Shack Color Computer and a video monitor.

Severe RFI from my radio transmitter into a television set used as a monitor convinced me to try a number of possible solutions. A low-pass filter at the transmitter output along with a high-pass filter at the input to the television set minimized the RFI somewhat, but not enough to call it a cure. I then decided to buy a black and white video monitor and use the direct video approach. Much better character resolution is also achieved by this method.

My first thought was to use a wide bandwidth operational amplifier with a gain setting high enough to produce a standard composite video amplitude of 1 volt peak-to-peak. The output signal from the op-amp is then applied to a black and white or color monitor. **Figure 8** shows an OEI model 9804 op-amp; however, any op-amp with a bandwidth of 30 MHz or higher should work well. Input to U1 is connected to U12, pin 12 in the TRS-80C. This point can be found at resistor R16 (3.9 k), located at the rear and center of the computer printed circuit board (component side up).

Connect the left-hand lead of the resistor to the input of U1 with miniature coax and ground the shield of the coax to the ground plane of the circuit board housing U1. Amplifier gain is set by adjusting R2. A simple alignment approach is to adjust the monitor contrast control for maximum contrast. R2 is then adjusted until the black and white picture has attained proper contrast and the sync pulses are not overloading the monitor. If a scope is available the op-amp output can be monitored until 1 volt peak-to-peak is observed.

A problem develops with the color signal from the computer driving a black and white monitor. The color burst signal riding on the composite video signal distorts the displayed characters observed on the screen. I solved the problem by installing a notch filter at the output of the op-amp tuned to the 3.58-MHz color burst frequency. While observing the monitor screen, adjust L1 until the characters look the best (you

should see sharp corners with no overshoot or smearing). Again, if a scope is available, L1 can be adjusted until the color burst disappears. The color burst will be seen as a burst of RF sinewaves after the horizontal sync pulse. If a color monitor is used the notch filter can be eliminated or switched out of the circuit as shown in **fig. 8**.

Approximately 40 mA is drawn from the op-amp so the power supply can be very simple in design. Any junkbox 12-volt center-tapped transformer can be used.

construction

I built the op-amp circuit on a perforated ground plane circuit board and enclosed it in a well-shielded aluminum box (very important for RFI elimination). The box should be thin enough to fit inside the computer enclosure underneath the top cover. I also installed the power supply inside the computer enclosure. The power supply parts were installed on the metal cover inside the computer (top side). Connect feed-through caps (1500 pF) onto the video circuit box in series with each power supply voltage lead. This will bypass the small amount of remaining RF from the transmitter into the video amplifier enclosure. Use miniature coax (RG174) between the output of the amplifier and the video monitor.

The results from this effort were very rewarding in that I can now run a kilowatt input from the transmitter with a clean screen. Character resolution was improved markedly because the maximum bandwidth of the monitor is realized. This is especially true with a black and white monitor such as mine, which has an 8 MHz bandwidth.

Don Kadish, W1OER

some thoughts on quad construction

Over the years, the quad antenna — frequently in the form of a three-band unit for 10, 15, and 20 meters — has been used extensively by Amateurs. It's a compact antenna that offers good performance.

spreaders

The quad's ability to survive strong winds depends largely upon the spreaders and how they're attached to the boom. The cast aluminum type of spider mounting on the boom is lightweight and relatively inexpensive, but may crack under wind stress. In an attempt to overcome this problem, I began, about 20 years ago, to use nylon monofilament as a means of reinforcing the spreaders (**fig. 9**). This method of reinforcement has been quite successful.

When the last major typhoon hit Manila in 1970, damage was extensive. Many houses, power and telephone utility poles, and main distribution lines were blown down. Most of the city was without power for two weeks. One of my antennas was a multi-band quad and delta system, with a total boom length of 50 feet, mounted atop a self-supporting 75-foot tower. The boom had a two-element 40-meter quad, a three-element 20-meter quad, a four-element 15-meter delta loop above the boom, and a five-element 10-meter delta loop below the boom. The largest antenna — which consisted of the 40-meter quad, the reflector of the 20-meter quad, and the director of the 20-meter quad — measured about 28 feet across. The rest of the boom was the bow sprit, where the antenna stays were attached. While one might not expect such a structure to survive a typhoon, the only significant damage was that the 2-inch solid steel shaft on which the boom was mounted was badly bent at the point at which it entered the retaining bearing at the top of the tower. The prop-pitch rotator, set about 5 feet lower, suffered no harm. There was a crack, however, in one of the 40-meter spreaders, which were made of the best material obtainable — phenolic resin molded with fiberglass.

Although I'd used 200-pound monofilament fishing line to reinforce the spreaders of this antenna, a less elaborate antenna, such as two- or three-element 20-meter quad, need not use anything heavier than line rated at 150 pounds. In such cases, the boom extension doesn't need to be quite as strong as the boom itself; the most practical thing to do is use the next smaller size of tubing — preferably about the same length as the quad spreaders themselves — so as to make a snug fit in the main boom. All assembly work should be done on the ground to make sure that all stays are of equal length, type, and shape. This means that the boom will have to be placed on an improvised sawhorse about 12 feet off the ground (for a 10-, 15-, and 20-meter antenna), so that the antenna and boom assembly can be rotated axially to provide access to all spreaders.

home-brewed spreaders

Hams in third-world countries may have difficulty finding spreaders; if they can be located, the cost of shipping may be prohibitive. Ordinary PVC pipe is now available almost anywhere, however, and can be used successfully.

There's a problem with PVC pipe, however. Even in the thicker wall sizes, it's too flexible and must be reinforced for rigidity. One way to do this is to obtain a piece of straight-grained lumber as long as the spreader and split it lengthwise into several square pieces whose dimension across the flats is slightly greater than the inner diameter of the tubing, using knives and coarse sandpaper to bring each to the correct dimension. One could also use a lathe.

After preparing the wood, it should be treated with preservative, allowed to dry thoroughly, and then inserted into the PVC pipe. Seal the ends with epoxy to keep moisture out. (It's a good idea to keep the wood slightly shorter than the pipe (perhaps one-fourth inch overall) to allow for this layer of epoxy.)

This type of spreader has been successfully used here in the Philippines. We find it will greatly outlast bamboo; it's also stronger. And while it may not compare with the best pressure-molded phenolic fiberglass, it costs much less. It does weigh more, but not as much as one might assume.

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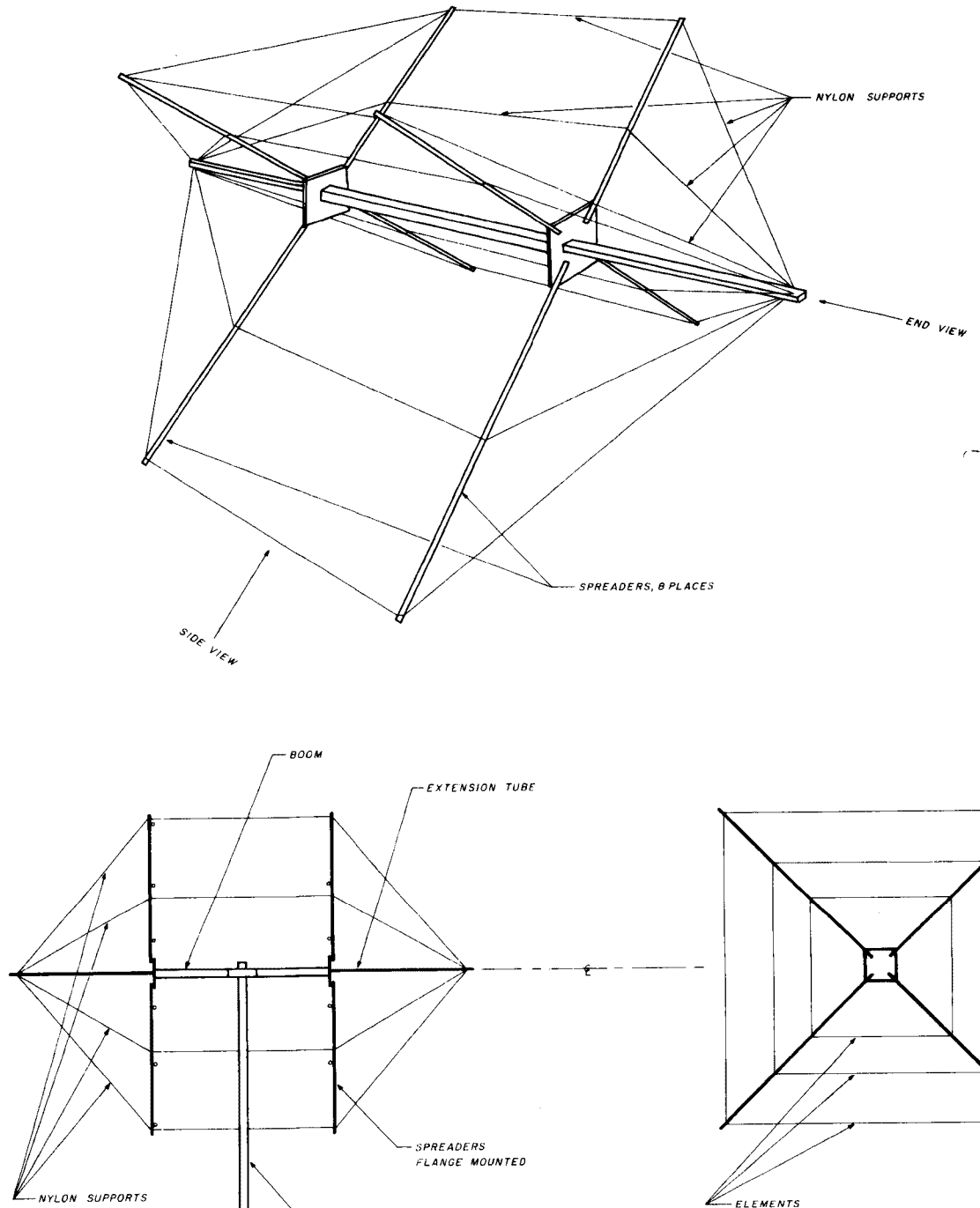


fig. 9. 10 through 20 meter quad antenna (A) three dimensional view shows spreader and nylon support structure; (B) end view (in the line of fire); (C) side view shows the relationship between the boom extension and elements.

selecting wire

Using insulated wire in a quad would result in unexpected difficulty in resonating the antenna. This is because the velocity of propagation is slower in insulated wire than in uninsulated wire in air.

It's possible to use magnet wire, which doesn't have this problem because its very thin coating has little effect. Older types of magnet wire may crack in the sun, but newer types, coated with tougher enamels, offer greater protection against corrosion and are more durable.

The ideal wire would be small-diameter copperweld such as 18 AWG, which has greater breaking strength than ordinary copper wire of larger diameter. I don't know of a source of enamel copperweld, but considering the advantages of copperweld — even without enamel — I would prefer to use it rather than any other material. Though it's more difficult to handle, less wind resistance and greater strength mean that there's less chance of damage from vibration in heavy winds.

Plated electric fence wire is even cheaper than copperweld and is strong and durable. It can be used on quads, although not in high-power applications.

weatherproofing

When putting up your coax, take particular care to protect the ends of the coax from moisture by using protective materials (for example, Coax Seal™). To make your antenna look more balanced to the transmission line, convert a section of the coax near the joint with the antenna, into a choke, winding the coax around a ferrite rod. The handbooks and various magazine articles have covered this subject; it costs little and will reduce susceptibility to noise with vertical polarization. If you use a two-element, three-band quad, you would probably find that a 75-ohm cable will give a better overall SWR than a 50-ohm cable, and for the same loss or power handling capacity, cost a bit less.

installation and adjustment

One of the serious problems of putting up a beam antenna is the difficulty in adjusting the antenna near the ground, whether it's a quad or Yagi. If one has a four-legged, self-supporting tower such as mine, the solution is easy — simply build a temporary scaffold (with a railing) extending out from the tower at the level necessary for working on the boom out to the farthest element. For other types of masts, you'll have to find your own solution, for the true tuning-up must be done on the spot.

Earl H. Hornbostel, DU1AE
Manila, Philippines

audio switcher for mobile operation

I don't know about you, but I have a very annoying problem. Even though blessed with two ears, I can't seem to really listen to more than one thing at a time. And, sad to say, the problem seems to be growing worse with age! Lately it seems that all the cocktail-party conversations and the rare DX seem to get mangled and mauled by QRM.

The problem was especially acute on the road while scanning the repeaters and trying to listen to the stereo at the same time. No amount of fiddling with the volume controls would produce the right result, and I was always lurching to turn down the stereo volume before pressing the PTT, which I usually squeezed instinctively. I'd be driving along the highway, shouting into the mic and groping for the stereo's volume control.

solving the problem

The circuit shown switches the car's speakers from the stereo to the rig whenever the PTT is pressed or whenever the rig's squelch is opened. Neither the rig nor the stereo has to be modified. Compact enough to fit into any small space, the circuit can be built and installed in a day. Parts selection is neither critical nor costly. Everything you need is probably right there in your junk box; even if you were to purchase all new parts, their cost should not exceed $15 to $20.

Before beginning construction, check to make sure that your stereo has four "hot" speaker leads and a common ground. Some stereo systems — Sanyo, for example — use a "matrixed" speaker output consisting of four speaker leads with no common ground. I know of no way around this except to use the phone or amplifier outputs and an external amplifier with the normal four "hot" channel outputs and one common ground.

The circuit shown in **fig. 10** consists of a quad op amp, two transistors, a relay, and a few resistors and capacitors. Construction techniques aren't particularly demanding. I pushed the parts through perf board and just soldered the component leads together (**fig. 11**). The

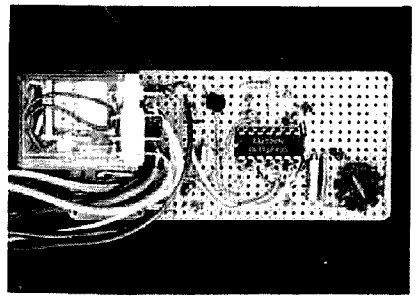


fig. 11 Simple perf board construction provides quick results.

relay was hot-glued to the perfboard and the completed unit was installed in a hinged plastic box that had once held standard-size staples. (A DIP socket was used for the op amp for development purposes, but it isn't necessary.)

operation

Audio output from the rig is fed into two successive stages of the LM324 op amp, where it's amplified and effectively clipped at a 12-VDC signal level. Capacitor C1 and variable resistor R1 determine the time constant for signal decay. This acts like a VOX delay to prevent relay dropout between words or pauses in the conversation. An additional op amp stage sharpens the signal and acts as a buffer. The signal is then fed into a 2N2222 PNP or equivalent transistor which drives the 4PDT relay. This relay switches all four speaker leads from the normal stereo position to the rig's audio output. The 1N4004 diode

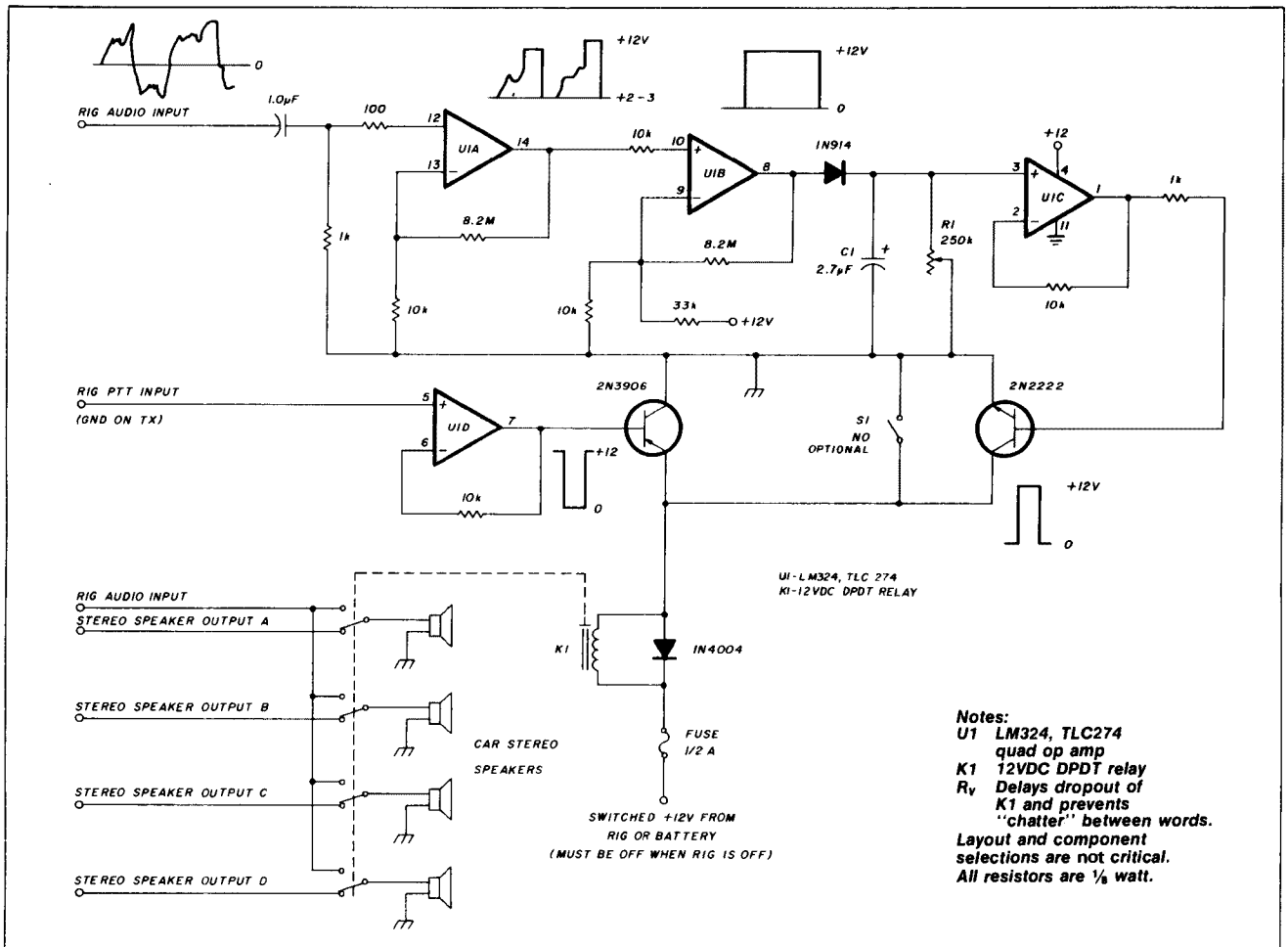


fig. 10. Audio switcher: layout and component choices are not critical.

shunts any harmful spikes when the relay opens. Sensitivity is high enough so that almost any audible output from the rig will operate the relay and switch the speakers from stereo to rig audio. The fourth op amp in the package is used to drive a 2N3906 NPN (or equivalent) transistor to pull in the relay when the PTT switch is pressed. This mutes the stereo system while you're talking into the microphone. Use of switch S1 is optional. It's normally used in the open position; when it's closed, it locks in the relay and the speakers are connected to the rig's audio output only. This may be useful under marginal receive conditions, but I usually just turn the rig off.

No tweaking or alignment is necessary after construction. However, you should double-check all wiring, solder joints, etc. If you've installed a socket for the op amp, leave it empty while you check relay operation and voltages. Check pin 4 for a nominal +12 VDC. Ground the transistor side of the relay coil and see if it actuates. If it's working up to this point, plug in the op amp and proceed. If not, double-check everything *again*.

Once everything is operating to your satisfaction, enclose it all in a small plastic case. At approximately 1 x 1.5 x 4 inches, my board fit into the staple box neatly, without movement. I placed a piece of non-conducting foam underneath the board and concealed the unit behind the ashtray above the radios. If space is a problem, you could, with a little effort, reduce the size of the board by at least 50 percent.

Operation is very simple. The only "set once and forget" adjustment is R1, which should be adjusted just as you would any VOX delay circuit. Listen to a few typical QSOs on the rig and adjust the resistance until the relay stops dropping out between words. That's all there is to it.

Some additional ideas came to mind after I built this nifty little circuit. One is to use some form of solid-state switching if relay noise becomes bothersome. Those running a "full house" of VHF, HF, scanners, and such may want to add a digital voting scheme. This would require a much costlier and complex circuit, however.

I've been using this circuit for several months. It's indeed a pleasure to listen to one radio at a time and to enjoy complete silence when keying the microphone button.

Emile F. Alline, Jr., NE5S

simple IC function generator

The **Exar XR-2206** is a widely available IC function generator (see **fig. 12**). The VCO frequency is controlled by the current drawn from either pin 7 or 8. Three volts is applied as bias on pins 7 and 8. Frequency of oscillation is given by:

$$f(Hz) = \frac{320 \cdot I\,(mA)}{C\,(\mu F)}$$

where *I* is the current drawn at pin 7 or 8 and *C* is the value of a capacitor placed across pins **5** and 6.

A typical method of varying the frequency is to place a potentiometer between pin 7 and ground. The frequency of oscillation now becomes

$$f = \frac{320 \cdot 3}{CR}$$

where R is the value of the pot. The frequency of oscillation is proportional to the inverse of the pot's resistance. If the pot has a linear taper, most of the frequency change takes place at the end of rotation. A better frequency control scheme is needed.

A solution is the voltage-to-current converter. A basic V to I converter is shown in **fig. 13**. The improved frequency control is shown in **fig. 14**.

The 2k turn pot has a maximum voltage range of about 2 volts and the V to I converter has a "gain" of 1V to 1mA

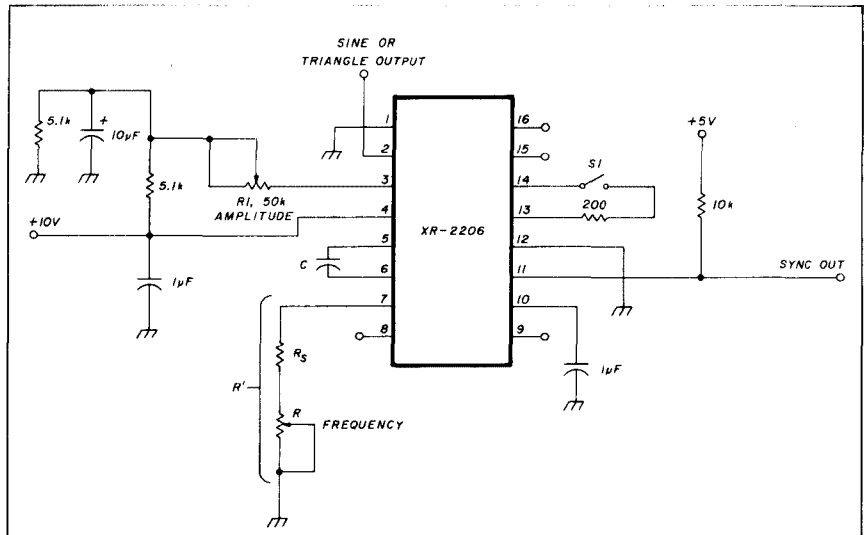


fig. 12. XR-2206 test circuit.

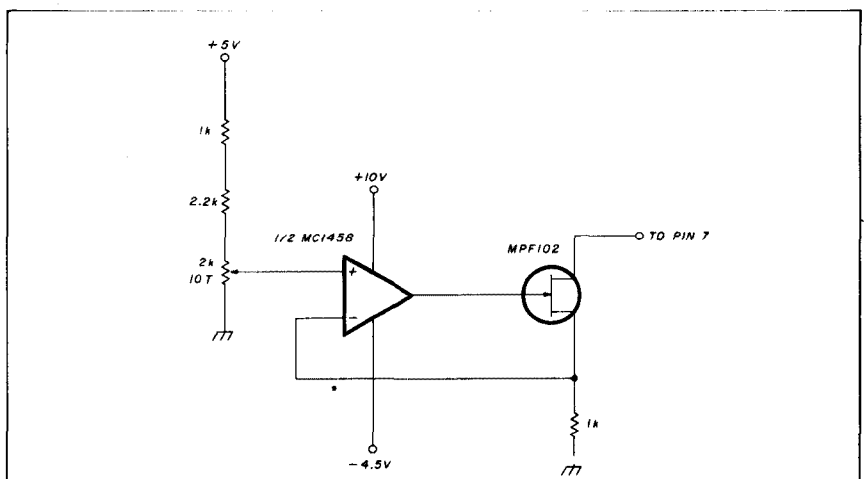


fig. 14. Improved frequency control.

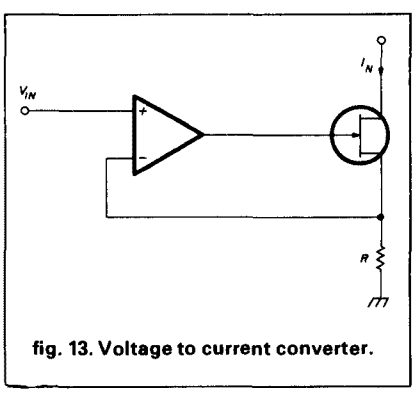


fig. 13. Voltage to current converter.

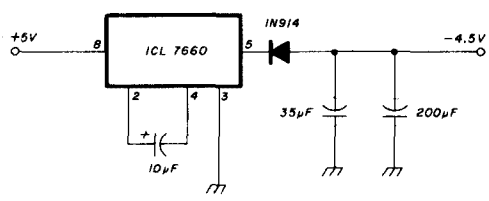


fig. 15. Regulated −4.5 volt supply.

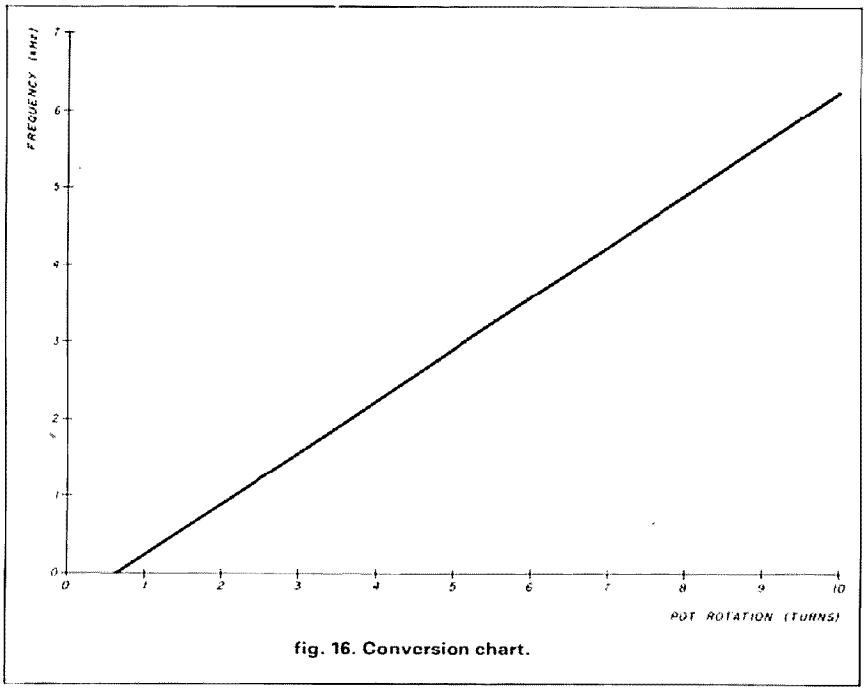


fig. 16. Conversion chart.

so that the maximum current drawn from pin 7 is about 2 mA. Exar says that current drawn from pin 7 or 8 should be less than 3 mA.

The Op amp is supplied with −4.5 volts to allow full-range operation down to 0V input. The −4.5 volts is supplied by an ICL 7660 voltage converter chip. The schematic is shown in **fig. 15**.

The 35 μF capacitor on the output provides extra filtering to reduce voltage variations to the V to I converter.

The results have been pleasing. **Figure 16** is a graph showing frequency versus pot rotation.

Thomas A. Keely

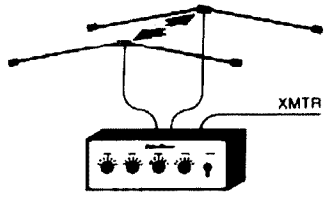
Yaesu FT707 DC power cable

Just a word of caution if you plan to make your own DC power cable for the Yaesu FT707 using off-the-shelf, four-contact Jones plugs. The pin numbering system of the Yaesu plug is different from that of the off-the-shelf plugs. On the Yaesu, pins 1 and 2 (horizontal blades) have a jumper, pin 3 is positive, and pin 4 is negative. On the off-the-shelf Jones plug, pins 1 and 3 should have the jumper, pin 2 is positive, and pin 4 is negative.

Jerome M. Havel, W2RRX

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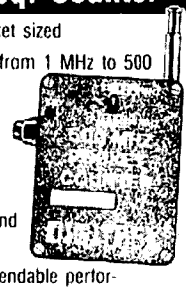
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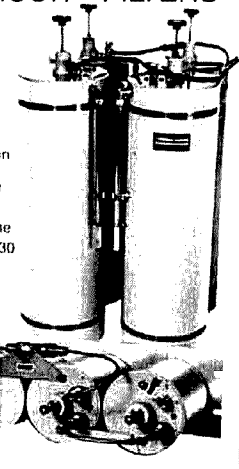
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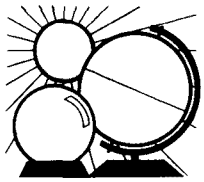


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Garth Stonehocker, KØRYW

antenna-to-ionospheric path match

Up North, October is probably your last chance to work on your antennas before winter and the contest season begin in earnest. Winds and temperature changes can take their toll. If your antennas and towers haven't been checked recently, hardware needs to be tightened and wires checked for cracks or worn places.

How have your antennas been performing? Perhaps a different antenna or configuration might be in order. At this point of the sunspot cycle, emphasis is now on the lower bands. What improvements could you incorporate to enhance your signal to your favorite DXing area?

Re-read K2RR's article, "Secrets of Successful Low-Band Operation (*ham radio,* May and June, 1986). Might there be something in it that would help improve your signal — analyzing your ground reflection areas, for example?

In January, 1985 I presented HORANT, a short computer program, written in BASIC, designed to provide the elevation angle, take-off angle, and the lobes of maximum radiation from a horizontal antenna mounted a selected height above ground for specific frequencies. I've since found that adding two lines to each of four places will also provide the ground reflection distance (associated with each lobe) away from the antenna feedpoint. A revised version of HORANT is listed in **fig. 1**.*

In part two of his article, K2RR pointed out (in **table 10**) that the quality of this reflection was the crux of the comparison of the efficiency of low horizontal antennas and verticals under poor ground conditions. Some horizontal antennas were able to provide more signal at the identical angle under poor soil conditions. HORANT can determine where your ground reflection point is and the take-off angle for comparison to the required radiation angle for your ionospheric path to your DX. (The latter angle can be found by using the modification of the MINIMUF 3.5 program or **table 1** of February, 1985's "DX Forecaster.")

After analyzing your ground reflection condition, there are a couple of things you can do to improve the situation. You can alter the reflection condition by soil management (add more radials or ground screen) or move the antenna to a point at which the ground reflection location is better. Relating the take-off angle and ground reflection terms is called *coupling the signal to the ionosphere*. Doing this allows for the efficient use of HF propagation, regardless of whether you're interested in conveying information over long distances or merely having some fun with DX.

last-minute forecast

Beginning with the first week of October and continuing through the second and third weeks of the month, MUFs are expected to be lower. This will result in poorer conditions on the 10- through 30-meter bands, making the lower bands — and nighttime conditions — more appropriate for DXing. During the fourth week, expect to see a return to better higher-band conditions until the

```
10  REM     PROGRAM HORANT, ANGLE & DISTANCE
20  R1=180/3.141659
30  PRINT "ANTENNA HEIGHT ABOVE GROUND IN FEET"
40  INPUT H1
50  PRINT "FREQUENCY IN MHz"
60  INPUT F1
70  REM   CALCULATE A FULL WAVELENGTH L1 AT THIS FREQ
80  L1=984/F1
90  REM CALCULATE FIRST LOBE ANGLE,A1
100 A1=ASN(L1/(4*H1))
110 D1=H1/TAN(A1)
120 A1=A1*R1
130 PRINT "ANGLES OF MAXIMUM RADIATION FROM THIS ANTENNA"
140 PRINT
150 PRINT "FIRST LOBE ANGLE,DEG=",A1
160 PRINT "DISTANCE TO GROUND REFLECTION IN FEET=",D1
170 PRINT
180 REM   CALCULATE 2ND LOBE ANGLE ,A2
190 IF (4*H1)<(3*L1) THEN 420
200 A2=ASN((3*L1)/(4*H1))
210 D2=H1/TAN(A2)
220 A2=A2*R1
230 PRINT "SECOND LOBE ANGLE IN DEG =",A2
240 PRINT "DISTANCE TO GROUND REFLECTION=",D2
250 PRINT
260 REM   CALCULATE THIRD LOBE ANGLE,A3
270 IF (4*H1)<(5*L1) THEN 420
280 A3=ASN((5*L1)/(4*H1))
290 D3=H1/TAN(A3)
300 A3=A3*R1
310 PRINT "THIRD LOBE ANGLE IN DEG=",A3
320 PRINT "DISTANCE TO GROUND REFLECTION=",D3
330 PRINT
340 REM   CALCULATE FOURTH LOBE ANGLE,A4
350 IF (4*H1)<(7*L1) THEN 420
360 A4=ASN((7*L1)/(4*H1))
370 D4=H1/TAN(A4)
380 A4=A4*R1
390 PRINT " FOURTH LOBE ANGLE IN DEG=",A4
400 PRINT "DISTANCE TO GROUND REFLECTION=",D4
410 PRINT
420 PRINT "NO(OTHER)SIGNIFICANT LOBES"
430 END
```

fig. 1. HORANT Program helps determine takeoff angles and distances to associated ground reflection points (written for IBM 4341 in IBM BASIC/VS).

WESTERN USA

GMT	PDT	N	NE	E	SE	S	SW	W	NW
0000	8:00	20	40	20	10	12	10	10	20
0100	8:00	20	*40*	20	12	12	10	10	20
0200	7:00	20	*40*	20	15	12	10	10	20
0300	8:00	30	*40*	20	15	15*	12	12	20
0400	8:00	40	*40*	*30*	20	15	15	15	40
0500	10:00	40	*40*	30	20	15	15	20*	40
0600	11:00	40	*40*	*30*	20	20	20	20	40
0700	12:00	40	*40*	*30*	20	20	20	20	40
0800	1:00	*40*	*40*	*30*	*20*	20	20	20	40
0900	2:00	*40*	*40*	*30*	*20*	*30*	20	20	40
1000	3:00	*40*	*40*	*30*	*30*	*30*	20	20	*40*
1100	4:00	*40*	*40*	*30*	*30*	*30*	*20*	*20*	*40*
1200	5:00	*40*	30	15	*30*	*30*	*30*	*30*	*40*
1300	0:00	*40*	20	12	15	*30*	*30*	*30*	*40*
1400	7:00	40	20	12	12	20	*30*	*30*	*40*
1500	8:00	40	20	10	12	20	*20*	20	*40*
1600	9:00	40	20	10	12	15	*20*	20	*40*
1700	10:00	*40*	20	10	10	15	20	20	*40*
1800	11:00	*40*	20	10	10	12	20	20	*40*
1900	12:00	*40*	20	12	10	12	15	20	20
2000	1:00	*40*	30	12	10	12	12	15	20
2100	2:00	*40*	40	15	10	12	12	12	20
2200	8:00	*40*	40	20*	10	12	12	12	20
2300	8:00	30	40	20	10	12	10	12	20

MID USA

MDT	N	NE	E	SE	S	SW	W	NW	CDT
0:00	40	40	20	15	12	10	10	30	7:00
7:00	40	40	20	15	12	10	10	40	8:00
8:00	40	*40*	20	20	15	12	12	40	9:00
9:00	40	*40*	20	20	15	15	15	40	10:00
10:00	40	*40*	*30*	20	20	15	15	40	11:00
11:00	*40*	*40*	*30*	20	20	20	20	40	12:00
12:00	*40*	*40*	*30*	*20*	20	20	20	40	1:00
1:00	*40*	*40*	*30*	*20*	20	20	20	40	2:00
2:00	*40*	*40*	*30*	*20*	*30*	20	20	*40*	3:00
3:00	30	*40*	*30*	*30*	*30*	*20*	20	*40*	4:00
4:00	20	30	20	*30*	*30*	*30*	*20*	*40*	5:00
5:00	30	20	12	*30*	*30*	*30*	*30*	*40*	6:00
6:00	30	20	12	20	*30*	*30*	*30*	*40*	7:00
7:00	30	20	10	15	30	*20*	*30*	40	8:00
8:00	40	20	10	15	20	*20*	*30*	40	9:00
9:00	40	20	10	12	20	20	20	*40*	10:00
10:00	*40*	20	10	12	15	20	20	*40*	11:00
11:00	*40*	20	10	10	15	20	20	*40*	12:00
12:00	*40*	20	10	10	12	15	20	30	1:00
1:00	*40*	20	12	10	12	15	20	30	2:00
2:00	*40*	30	15	10	12	12	15	20	3:00
3:00	*40*	40	15	10	12	12	12	20	4:00
4:00	*40*	40	20	12	12	12	12	20	5:00
5:00	30	40	20	15	15	10	12	20	8:00

EASTERN USA

EDT	N	NE	E	SE	S	SW	W	NW
8:00	40	40	20	12	12	12	12	30
9:00	40	40	20	15	15	20	20	30
10:00	40	*40*	20	20	15	20	20	40
11:00	40	*40*	20	20	20	20	20	40
12:00	*40*	*40*	*30*	20	20	20	20	40
1:00	*40*	*40*	*30*	20	20	20	20	40
2:00	*40*	*40*	*30*	*20*	20	20	20	40
3:00	*40*	*40*	*30*	*20*	*30*	*30*	*30*	*40*
4:00	*40*	*40*	*30*	*20*	*30*	*30*	*30*	*40*
5:00	*40*	30	20	*30*	*30*	*30*	*30*	*40*
6:00	40	20	12	*20*	*30*	*30*	*30*	*40*
7:00	20	20	12	20	30	*30*	*30*	40
8:00	20	20	10	15	20	20	20	40
9:00	30	20	10	15	20	20	20	40
10:00	40	20	10	12	15	20*	20	40
11:00	40	20	10	12	15	30*	20	*40*
12:00	40	20	10	10	15	20	20	*40*
1:00	40	20	10	10	12	20	20	*40*
2:00	*40*	20	10	10	12	15*	20	*40*
3:00	*40*	30	12*	10	12	15	20	30
4:00	*40*	30	12	10	12	12	15	20
5:00	*40*	40	15	10	12	12	12	20
6:00	*40*	40	20*	10	12	12	12	20
7:00	*40*	40	20	12*	12	12	10	20

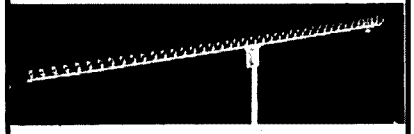
OCTOBER

	N	NE	E	SE	S	SW	W	NW
ASIA FAR EAST	0000	0000	0000	0000	0000	0000	0000	0000
EUROPE	0000	0000	0000	0000	0000	0000	0000	0000
S. AFRICA	0000	0000	0000	0000	0000	0000	0000	0000
S. AMERICA	0000	0000	0000	0000	0000	0000	0000	0000
ANTARCTICA	0000	0000	0000	0000	0000	0000	0000	0000
NEW ZEALAND	0000	0000	0000	0000	0000	0000	0000	0000
OCEANIA AUSTRALIA	0000	0000	0000	0000	0000	0000	0000	0000
JAPAN	0000	0000	0000	0000	0000	0000	0000	0000

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.

*Look at next higher band for possible openings.

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end of the month; transequatorial openings associated with a disturbed geomagnetic field from the 20th to 23rd may even be possible. Another period of disturbance may occur from the 5th to 7th. Because both of these disturbances are within the equinox season, they may have more effect than we've seen recently. Look for openings on east-west paths, with DX signals arriving from some unusual directions.

The Orionids meteor shower will be visible from the 15th to 24th of October, with a maximum rate of between 10 to 20 per hour on the 20th and 21st of the month. A full moon will occur on the 25th; perigee will take place on the 7th. An annular eclipse of the sun extending from Alaska to northern South America begins on October 3 at 1657 UTC. It moves to the east, reaching Eastern Canada and Greenland by 2030 UTC, then ending at 2114 as it approaches Great Britain and Africa.

band-by-band summary

Ten, twelve, fifteen, and twenty meters provide many openings during the daytime. The higher bands will experience shorter openings; these will occur around local noon mainly in a southerly direction. Fifteen meters is now only a transition band between 12 and 20. Twenty is the daytime band for the northern directions and transequatorial openings may occur in the evening. Distances may reach 2000 miles per hop if antenna take-off angles are as low as 10 degrees.

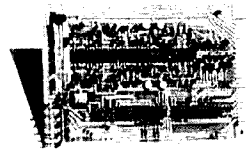
Thirty, forty, eighty, and one-sixty meters are all useful for nighttime DX. Thirty and 40 meters are the night frequencies for the east-west and northern directions and for distances of up to 1600 miles tivity. In these days of low solar activity, the MUFs can drop as low as the 80-meter band frequencies, resulting in higher propagated signal levels. These bands should be much quieter now that the thunderstorms have moved back down to the tropics.

*To run this program on the Apple IIe or IIc, or on other machines that don't have an intrinsic ARC SIN (ASN) function, substitute the following lines. – Ed.

```
 95  IF 4*H1 < L1 THEN 420
 98  Y = (L1/(4*H1))
100  A1 = ATN (Y/(1-Y 2) 0.5)
198  Y = (3*L1)/(4*H1)
200  A2 = ATN (Y/(1-Y 2) 0.5)
278  Y = (5*L1)/(4*H1)
280  A3 = ATN (Y/(1-Y 2) 0.5)
358  Y = (7*L1)/(4*H1)
360  A4 = ATN (Y/(1-Y 2) 0.5)
```

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MMT1296/144G	1296-1298MHz, 144at, 2w, GAASFET	$369
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software for ICOM repeaters

Processor Concepts has announced its new RPS-1 repeater program for the ICOM repeater series. The RPS-1 software is a program written for the ICOM repeater series that will add new features and custom programming to ICOM RP-1210s, RP-2010s, RP-3010s, or commercial repeaters.

A direct EPROM replacement for the ICOM repeater controller memory chip, the RPS-1 supports many new features such as two separate CW-ID messages and enough memory to include a club name or city and state. The CW-ID is sent at the speed you request; the ID timer is programmable from 30 seconds to 10 minutes. The timeout timer can be reset without waiting for the repeater transmitter to drop out. A courtesy beep tone signals the next user; the repeater stays on the air for a smooth transition. Pre- and post-timeout CW warning messages keep users informed and reduce confusion during timeouts.

The software is custom-programmed for individual CW and timing requirements. A reprogramming service is also available.

The RPS-1 software is priced at $74.00 plus $3.00 shipping. For information, contact Processor Concepts, P.O. Box 26023, St. Paul, Minnesota 55126.

Circle #301 on Reader Service Card.

half-wave VHF Kulduckie™

Larsen Electronics has announced a new half-wave Kulduckie™ portable antenna, Model KD14-2M-HW. The VHF antenna operates at frequencies from 144-148 MHz with a resonant half-wave design that allows it to function independently of a ground plane. Its performance is said to equal that of a full quarter-wave on a perfect ground plane. The rigid impedance transformer at the base contributes to performance and adds extra strength.

Finished in black chrome, the KD14-2M-HW also telescopes, using internally shorted collapsing joints. The unit extends to 41 inches for operation and collapses to 7-3/4 inches. The antenna uses a special double spring BNC connector to reduce wear in the female contact on the radio.

For details, contact Larsen Electronics, P.O. Box 1799, Vancouver, Washington 98668.

Circle #302 on Reader Service Card.

tone panel offers digital CTCSS

Digital CTCSS is now available as an option on Communications Specialists' TP 38 Shared Repeater Tone Panel. The new option, called the

TP-DCS, is compatible with Motorola's Digital Private Line,™ General Electric's Digital Channel Guard,™ and E.F. Johnson's Digital Call Guard.™ A TP-38 equipped with TP-DCS allows up to 14 DCTCSS subscribers as well as 37 CTCSS subscribers in a single repeater panel. Tone translation may be made from one DCTCSS code to any other DCTCSS code. Time and hit accumulation, remote access, and remote data retreival function are applicable to all DCTCSS subscribers. All DCTCSS codes between 000 and 777 Octal are available in normal or inverted polarity; squelch tail elimination is provided.

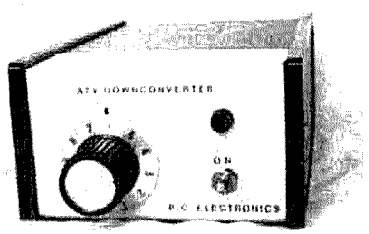
The TP-DCS is available as an option in new TP-38 Shared Repeater Panels or may be retrofitted into existing TP-38s. Priced at $149.95, the TP-DCS is in stock for immediate shipment and is covered by a one-year warranty.

For information and a free catalog, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665-4296.

Circle #303 on Reader Service Card.

GaAsFET ATV downconverter

A new GaAsFET ATV downconverter for the new 33-cm (902-928 MHz) band from P.C. Electronics features a dual-gate GaAsFET in both the RF preamp and the mixer stage for low noise and wide dynamic range. A varicap VCO tunes the whole band down to TV channel 3 (2 or 4 can also be used). Total conversion gain is + 25 dB, more than enough to reach the noise floor with most TV sets (even if one wants to remote the downconverter at the antenna to get the most sensitivity.)



The new band will provide relief for ATVers in high population density areas plagued with interference from radar, geophysical survey locators, and other mode users in the 70-cm (420-450 MHz) band. Parts and equipment should be more available and less expensive in this band than in the 12-cm band because of the proximity of cellular and business-band radios. Path loss is a little less too.

Model TVC-9G includes a wall-mounted 120-VAC to 12-VDC power supply. Just add an antenna, coax, and your TV set. Antenna input is a BNC connector, and the TV output is type F. The shielded cabinet measures 4 X 2.5 X 7 inches.

For those who want to make their own cabinet, the TVC-9 is available as a wired and tested board measuring 2 X 4 inches. With the on-board regu-

The HF4B "Butterfly"™

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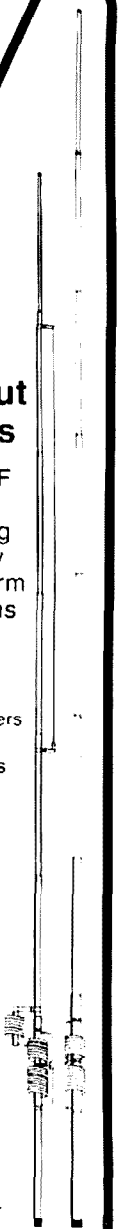
Model HF6V

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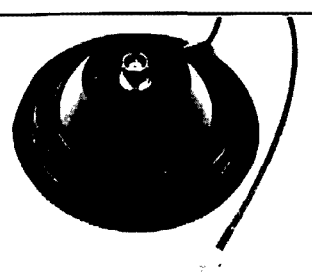
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For HT owners operating inside a vehicle and wanting increased T/R range, RF PRODUCTS has the low cost solution.

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The magnet mount (part no. 199-445) has 10 feet of small (5/32") co-ax with BNC conector attached. PRICE $15.95 M.O. or cashiers ck., via UPS gnd. Fla. residents add 5% tax, for air UPS add $3.25

The RF PRODUCTS Magnet Mounts are one of the few mounts available that can be repaired should the co-ax cable be damaged. The large surface area capacitance disc provides proper ground plane coupling for 1/4 and 5/8 wavelength VHF and UHF antennas.
MODELS AVAILABLE WITH THE FOLLOWING CONNECTORS & CO-AX TYPES.
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RF PRODUCTS

P.O. Box 33, Rockledge, FL 32955, U.S.A. (305) 631-0775

lator, the power requirement can be anywhere from 11 to 18 VDC at 30 mA. A removable 10-k trim pot is used for frequency tuning.

P. C. Electronics also offers a DC decoupler (Model DCD) and remote control box (Model DCBI for tuning and powering up through the channel-3 downconverter output.

For more information, contact P.C. Electronics, 2522 Paxson Lane, Arcadia, California 91006.

Circle #304 on Reader Service Card.

compact mobile radio mounts

IIX Equipment, Ltd. has announced the addition of two new compact mobile radio mounts to its Mount 100 system. The additions bring the model count to four, which can handle all types of communications gear needed in mobile operations. Models MM1007 (7 inches) and MM10010 (10 inches) represent a new design that provides the user with a strong, adjustable two-radio mount that will fit in most vehicles. The mounts are constructed of welded steel finished in satin black, with necessary hardware supplied. UPS shipping is included in the price of **$79.50** for the smaller MM1007 and **$84.50** for the larger MM10010. A free catalog and information sheet are available from IIX Equipment, Ltd., P.O. Box 9, Oaklawn, Illinois 60454.

Circle #305 on Reader Service Card.

coax lightning arresters

New coax lightning arresters from Cushcraft feature fast-action gas discharge elements. Four models offer choices of 200-watt or 2-kw power handling and type N or UHF connectors. They're precisely manufactured to give very low insertion loss from 2 to 1000 MHz. Replacement cartridges are available. Dimensions are 3.5 x 1.375 x 1.875 inch.

For more information, contact Cushcraft Corporation, P.O. Box 4680, Manchester, New Hampshire 03108.

Circle #306 on Reader Service Card.

digital capacitance tester

Mercer Electronics, a division of Simpson Electric Company, has introduced a new digital capacitance tester. The Model 9670 features nine ranges that will measure from 0.1 pF to 20,000 μF with 0.5 percent accuracy.

Priced at $99, the Model 9670 has a 0.5-inch LCD display with over-range and an "LO BAT" indication. The tester measures 6.85 x 3.54 x 1.42 inches, weighs 3/4 pound and uses a standard 9-volt battery. Convenient features include input discharge protection, easy-insert "cap-lead" jacks, color-coded test leads with alligator clips, a flame-retardant plastic case with an acrylic window, and a tilt bail.

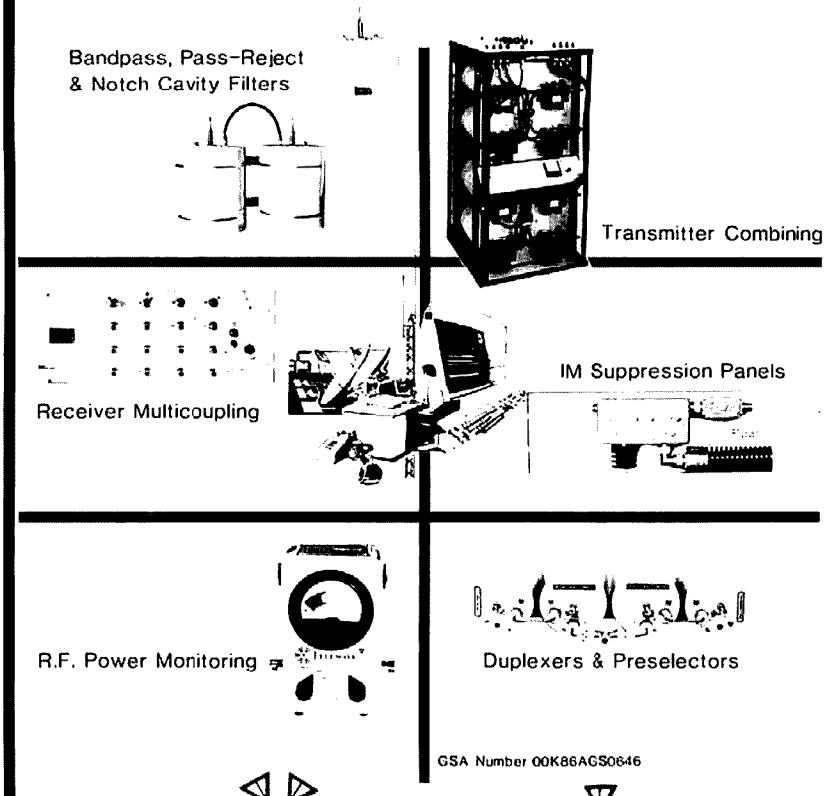
For information write Mercer Electronics, 859 Dundee Avenue, Elgin, Illinois 60120.

Circle #307 on Reader Service Card.

(continued on page 106)

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ANTIQUE RADIOS, schematics, tubes and literature. Send SASE to VRS(HR), 376 Cilley Rd, Manchester, NH 03103 for large list.

SPEECH SYNTHESIZER KIT FOR C-64 uses allophones to obtain virtually unlimited vocabulary. P.C. board, all parts, instructions $30.00 plus $2.00 UPS. Dave Roscoe, W1DWZ, 49 Cedar Street, East Bridgewater, MA 02333.

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$$$$$SUPER SAVINGS on electronic parts, components, supplies and computer accessories. Free 40-page catalog for SASE. Get on our mailing list. BCD ELECTRO, PO Box 830119, Richardson, TX 75083 or call (214) 690 1102.

R390A/URR Rcvr .5-32 MHz, Central 20A Trx w/458 vfo, all real good condition. $135.00. 1120 VAC, 500 ADC transformer $15.00. U/pay shipping. Mike Nichols, KE0AD (303) 665-5104.

MEASUREMENTS CORP. signal generators, etc. Copies of instructions. Send SASE and Model for data. W2BLL, RD2, Box 72, Boonton, NJ 07005.

TALKING DIGITAL DIAL for ICOM 720A, very useful for partially sighted or blind Amateurs. Uses Kenwood speech brd. Plugs into 24 pin acc. socket. $125.00. Also a frequency counter version for almost any transceiver. $300.00. Jim Upright, 707 Huntridge Hill NE, Calgary, Alberta, Canada T2K 4A4. Ph. (403) 274-4284.

DYNAMIC COLOR NEWS: Monthly newsletter provides programming instruction for Radio Shack Color Computers. Includes HAM RADIO section. FREE sample. Dynamic Electronics, Box 896, Hartselle, AL 35640. (205) 773 2758.

CABLE TV CONVERTERS & EQUIPMENT: Plans and parts. Build or buy. SASE for information. C & D Electronics, PO Box 1402, Dept. HR, Hope, AR 71801

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R-390A Receiver: $195 checked; $115 reparable. Parts, tubes, sections. Info SASE. CPRC-26 six meter transceiver (see HR, March 1985) $17.50 apiece, $32.50 pair (add $4.50/unit shipping). Baytronics, Box 591, Sandusky, OH 44870. 419-627-0460 evenings.

SOME QSLs 100 for $6.25. Sample SASE. WB2EUF, Box 708, East Hampton, NY 11937.

MARCO: Medical Amateur Radio Council, Ltd. operates daily and Sunday nets. Medically oriented Amateurs (physicians, dentists, veterinarians, nurses, physiotherapists, lab technicians, etc.) invited to join. Presently over 550 members. For information write MARCO, Box 73's, Acme, PA 15610.

IBM-PC RTTY/CW. New CompRtty II is the complete RTTY/CW program for IBM-PC's and compatibles. Now with larger buffers, better support for packet units, pictures, much more. Virtually any speed ASCII, BAUDOT, CW. Text entry via built in screen editor! Adjustable split screen display. Instant mode/speed change. Hardcopy, diskcopy, break-in buffer, select calling, text file transfer, customizable full screen logging. 24 programmable 1000 character messages. Ideal for MARS and traffic handling. Requires 256k PC or AT compatible, serial port, RS-232C TU $65. Send call letters (including MARS) with order. David A. Rice, KC2HO, 25 Village View Bluff, Ballston Lake, NY 12019.

ELECTRONIC ENCLOSURES: Attractive, reasonably priced Aluminum top and sides. Walnut stained solid wood ends. Variety of sizes and custom sizes available. Sample $5.00 Information free. Energy Engineering, Custom Division, Rt. 4, Box 330, Fayetteville, AR 72701.

NATIONAL RADIO EQUIPMENT manual list or NCL-2000 parts kits. SASE. Maximilian Fuchs, 11 Plymouth Lane, Swampscott, MA 01907.

ATTENTION AMATEURS Send for Free discount catalog. Amateur Communications, 2317 Vance Jackson, San Antonio, TX 78213. (513) 734-7793.

8877 VHF AMP KITS: HV power supplies, CX600N relays, Mu Tek LTD front end boards for IC251/IC271, EME newsletter and QRO parts. SASE for new catalog. KB7Q, "Q" Products, 417 Staudaher Street, Bozeman, MT 59715.

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THE GOOD SAM HAMS invite RV operators to check in the Good Sam Ham net 14.240 *Sundays 1900Z also 3.880 Tuesdays at 2359Z Net control N5BDN, Clarksville, IN*

IMRA, International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, New York 10538.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

DISCOUNT CATV CONVERTERS/DECODERS and Video accessories. Send for free information and prices - it could save you BIG MONEY on your next purchase of these and other CATV items. Easy View, (HR) PO Box 221K, Arlington Heights, Illinois 60006. (312) 952-8504. Ask for Rudy Valentine.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. Daily Electronics, PO Box 5029, Compton, CA 90224. (213) 774 1255.

CUSTOM MADE EMBROIDERED PATCHES. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept 301, 4202 N. Drake, Chicago, IL 60618.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

CABLE TV CONVERTERS/DESCRAMBLERS. Guaranteed lowest prices in US. Jerrold, Hamlin, Zenith ... Many others. Lowest dealer prices! Orders shipped within 24 hours! Master card, VISA, COD accepted. Free Catalog - call (800) 345-8927 only Pacific Cable Co., Inc., 7325-1/2 Reseda Blvd,#1017, Reseda, CA 91335. (818) 716-5914.

CHASSIS and cabinet kits. SASE K3IWK, 5120 Harmony Grove Road, Dover, PA 17315.

COMING EVENTS

Activities — "Places to go . . ."

OKLAHOMA: October 5. The Salt Plains Amateur Radio Club will hold their annual Ham Social, Salt Plains Lake, Northern Alfalfa County. Covered dish dinner. Talk in on 147.30/90. Inquiries to: Gary Gerber, KB0HH, 511 Lincoln, Anthony, KS 67003. (405) 842-5076.

MARYLAND: October 5. The Columbia Amateur Radio Association's 10th annual Hamfest, Howard County Fairgrounds, just west of Baltimore, off I-70. 8 AM to 3:30 PM. Admission $3.00. Spouse and children free. Reserved tables $7.00 prior to September 30. $8.00 after. Outdoor tailgating $3.00 additional. Indoor tailgating $6.00 additional. Food available. Talk in on 147.735/135, 146.52/52. For table reservation and information: Mike Vore, W3CCV, 9098 Lambskin Lane, Columbia, MD 21045. 992-4953.

PENNSYLVANIA: October 4. Pack Rats (Mt. Airy VHF ARC) invites all Amateurs to the 10th annual Mid-Atlantic VHF Conference, Warrington Motor Lodge, Rt 611, Warrington and the 15th annual Pack Rat Hamarama, Sunday, October 5. Bucks County Drive-In theater, Rt. 611, Warrington. Flea market admission $5.00 per car. Selling spaces $6.00 each. Gates open 6 AM rain or shine.

GEORGIA: November 1-2. The Alford Memorial Radio Club of Stone Mountain is sponsoring Ham Radio and Computer Expo '86, Gwinnett County Fairgrounds, 20 minutes NE of Atlanta. 9 AM to 5 PM Saturday; 9 AM to 4 PM Sunday. Admission $4

NEW products

new mic

Astatic has confirmed the rumor that they're working on a new microphone. The new model, titled the D-204, is a scaled-down version of the D-104 with new features. Said to have the same high-luster chrome finish as the Silver Eagle D-104, it will carry the trade name "Silver Classic." VOX compatibility and an adjustable preprocessor are some of the features expected to be included. For information, contact P.O. Box 120, Conneaut, Ohio 44030-0210.

Circle #310 on Reader Service Card.

produce prototype circuits direct from CAD or artwork

Girard Electronics has developed a non-chemical system for generating flexible prototype circuits which the user can laminate into double-sided PCBs, multilayer circuits and boards for standard or surface-mount components.

Protoflex-III uses optical-mechanical technology to produce prototype circuits. An optical system scans circuit artwork mounted on the right side of a drum within the machine. Protoflex-III machines circuits from 1 oz. copper sheeting on a polymide base affixed to the left side of the drum.

Protoflex-III works from a variety of artwork media — transparent, translucent or opaque. The user can produce artwork simply with graph paper and taping puppets or use camera shots or CAD plots, positives or negatives.

Additionally, the machine comes with CAD interface that allows the circuit designer to proceed directly from schematic capture to printed board.

Protoflex-III handles materials as large as 7 x 16 inches and machines flexible circuits in 140 minutes or less, depending upon the size of the board, with resolving power down to 0.010 inches in lines and spaces.

The system's technique eliminates messy chemicals used in acid etching systems and allows the machine to be housed in a desktop-sized case measuring approximately 31 x 10 x 17 inches and weighing about 70 lbs.

For information, contact Girard Electronics, 13914 West Oakgreen Circle South, Afton, Min-

advance, $5 at door. Forums, awards, VEC exams both days, free cookout Saturday night. Activities for the entire family. Superb dealer facilities, giant undercover flea market. Discount hotel rooms. Free parking, RV sites with full hookups. Talk in on 146.16/76, 449.25/4.25. Information: Alford Memorial ARC, PO Box 1282, Stone Mountain, GA 30086 or call N8LM at 404 925-7615.

NEW HAMPSHIRE: October 4. The Hosstraders Fall Tailgate Swapfest, Deerfield Fairgrounds. Donation $2 per person. Friday night camping at nominal fee. Entry 4 PM. Profits benefit Shriner's Hospital. Last Spring's gift was $7767.00. Sponsored by W1GWU, K1RQG and WA1IVB. For map SASE to WA1IVB, RFD Box 57, West Baldwin, ME 04091.

NEW HAMPSHIRE: October 4. In conjunction with the Hosstraders Swapfest, the volunteer examiners of NH will conduct exams at the Northwood Elementary School Gymnasium, Deerfield, NH. For further information SASE to Don Clark, N1AKS, 2 Cortland Street, Londonderry, NH 03053.

OHIO: October 26. The Marion Amateur Radio Club will hold its 12th annual "Heart of Ohio" Ham Fiesta. 0800 to 1600 hours, Marion County Fairgrounds Coliseum. Tickets $3.00 advance, $4.00 at door. Tables $5.00. Check in on 146.52 or 147.90/30. For information, tickets or tables contact Ed Margraff, KD8OC, 1989 Weiss Avenue, Marion, OH 43302. (614) 382-2608.

CONNECTICUT: November 9. SCARA annual indoor Flea Market, North Haven Park and Recreation Center, North Haven. Sellers 7 AM. Buyers 9 AM to 3 PM. Tables $10.00 advance, $15.00 at the door. Buyers $2.00. For information or table reservations SASE with phone number to SCARA Flea Market, PO Box 81, North Haven, CT 06473. Reservations must be received by November 3, 1986. No reservations taken by phone. For information ONLY contact Brad (203) 265-6478, 7 PM to 10 PM.

KANSAS: October 11-12. ARRL Convention. Holiday Inn Plaza, 250 W. Douglas, Wichita, KS 67202. Starts 9 AM both days. Indoor flea market, OCWA Meeting, ARRL Forum. FCC Booth, Ladies' lounge. VE exams, EME/OSCAR basics. Banquet with entertainment. Sunday buffet breakfast. Hospitality suite. Kansas Amateur of the Year award. Talk in on W0KA repeater, 146.82. For further information: Kansas ARRL Convention, 1520 W. 16th, Wichita, KS 67203.

MASSACHUSETTS: October 26. The Framingham Amateur Radio Association will hold its annual Fall Flea Market and Exams, Framingham Civic League Bldg, 214 Concord Street, Downtown Framingham. Doors open 10 AM. Sellers setup 8:30 AM. Admission $2.00. Tables $10.00, includes one free admission. Preregistration for tables and exams required. Talk in on 75/15 repeater. To reserve tables: Jon Weiner, K1VVC, 52 Overlook Drive, Framingham, MA 01701. (617) 877-7166. To register for exams send Form 610, copy of ham license and check for $4.25 payable to ARL/VEC to FARA, PO Box 3005, Framingham, MA 01701.

ALABAMA: October 25 and 26. The Montgomery Amateur Radio Club's 9th annual Hamfest and South Net Convention, Civic Center, downtown Montgomery. 9 AM to 4 PM. Free admission, free parking, all indoors, including Flea Market. Flea Market reservations not required. Setup begins 6 AM. FCC exams both days. Bring copy of current and $2.00. For more information: Hamfest Committee, 2141 Edinburgh Drive, Montgomery, AL 36116 or phone Phil (205) 272-7980 after 5 PM CST.

MASSACHUSETTS: November 8. New England annual DXCC Banquet, Masonic Lodge Building, Monument Square, in historic Concord Center. Afternoon session 2 PM to 5:30 PM. Banquet 6:30 PM. For further information: Charles Lukas, Jr., W1DOH, RFD 1, 24 Durkee Road, Acton, MA 01720. (617) 263-3743.

CALIFORNIA: October 4. Scatcon '86. 9 AM to 3 PM, Cortez Park, 2441 Cortez Avenue, West Covina. Tech sessions. Hands-on packet, satellite and more. Swaps, vendors. Donation $2. Talk in on 147.765-600. For more information: Bob, N6NGN. (818) 917-6470.

FLORIDA: October 18, 19. ARRL South Florida Section Suncoast Convention. Sponsored by the Florida Gulf Coast ARC National Guard Armory, St. Petersburg. Admission $4.00 advance; $5.00 at the door. 9 AM to 4 PM both days. QCWA luncheon Saturday. Western Barbeque Saturday night. Swap tables $12.00 for both days. Plenty of free parking. Some RV spaces. For information: FGCARC, PO Box 157, Clearwater, FL 33517.

MINNSOTA: November 1. The 2nd annual Hamfest Minnesota and Computer Expo, sponsored by the Twin City FM Club, Richfield High School, 7001 Harriet Avenue South, Richfield. 7:30 AM to 3 PM. Admission $3.00 advance, $4.00 at the door. Wayne Green, W2NSD, guest speaker. DX seminar, FCC exams, indoor flea market, ARRL, AMSAT and computer seminars and much more. Talk in on 16/76. For more information, registration SASE to Hamfest Minnesota and Computer Expo, Box 555, Minneapolis, MN 55440 or Lyly Vogt, KA0UDL, 5130 Willow Lane, Minnetonka, MN 55345.

PENNSYLVANIA: November 2. The RF Hill ARC's 11th annual Hamfest, Pennsylvania National Guard Armory, Rt. 152, Sellersville. Dealer setup 6 AM. Buyers 8 AM. Admission: Hams $4.00. Accompanying non-ham spouse and children free. Dealers 8x6' inside space $8.00, $6.00 outside. Own tables. Talk in on 144.71/145.31 Almont, PA, 146.28/146.88 Hilltown, 146.52 simplex. To reserve space: Frank Benner, W3BRU (215) 257-2450 or write Hamfest Chairman, RF Hill ARC, 523 Vine Street, Perkasie, PA 18944.

MICHIGAN: November 2. The Oak Park ARC will hold its largest ever 1986 SWAP-N-SHOP. New location: City of Southfield Civic Pavilion, Evergreen Road between 10 and 11 Mile Roads, northwest Detroit suburb of Southfield. Ham Radio and Computer activities from 9 AM to 5 PM in the new fully carpeted 30,000 sq. ft. pavilion. The Detroit Area Repeater Team (DART) will provide food and refreshments. Admission $4, under 12 free. VE3's at par. Tables $10.00. Advanced reservations required. Talk in on DART 146.04/.64 and 146.52 simplex. For further information SASE to OPARC Swap-N-Shop, 303 South Vermont Avenue, Royal Oak, MI 48067. Swap-N-Shop Hotline (313) 399-3991.

TENNESSEE: October 11, 12. The Mid-South Amateur radio Association and allied clubs will hold their annual Hamfest, Shelby County Agricenter (New Location), Memphis. Saturday 8 AM - 5 PM. Sunday 9 AM to 4 PM. Dealers and flea market inside large air-conditioned building. Flea market spaces $6.00 each per day. Tables furnished. Non-hams and children admitted free. Setup time Friday, 5 PM to 9 PM. Doors open Saturday morning for vendors/dealers 6 AM. FCC exams for all license classes. Bring license and copy, and Form 610. Hospitality party Friday evening at Hamfest Motel, Admiral Benbow Inn, free to all hams. Talk in on 28/88. Alternate 22/82. For further information: Clayton Elam, K4FZJ, 28 No. Cooper, Memphis, TN. (902) 274-4418 days. (901) 743-6714 nights.

1986 "BLOSSOMLAND BLAST" Sunday, October 5, 1986. Write "BLAST", PO Box 175, St. Joseph, MI 49085.

CALIFORNIA: FCC exams, Novice-Extra. Sunnyvale VEC ARC. (408) 255-9000 24 hour. 73, Gordon, W6NLG, VEC

MASSACHUSETTS: The MIT UHF Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday, October 22, 7 PM, MIT Room 1-134, 77 Mass Avenue, Cambridge MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-0160/646-1641 or Craige Rodgers (617) 494-1986. Exam fee $4.25. Bring copy of current license (if any), two forms of picture ID and completed form 610 available from FCC in Boston (223-6609).

NEW YORK: October 5. Electronics Fair and Giant Flea Market, Yonkers Municipal Parking Garage, Nepperhan Avenue and New Main Street. 9 AM to 4 PM. Satellite TV, SSTV, Amateur Radio, Computers and more. Giant Auction 2 PM. Unlimited free coffee all day. Admission $3.00. Children under 12 free. Sellers $7.00 per parking space, one admission. Bring tables. For information call (914) 969-1053.

MICHIGAN: October 12. Ham Fair '86, Michigan National Guard Armory, 2500 S. Washington Avenue, Lansing. 8 AM to 3 PM. Dealer sales, swap shop, handcrafted items, FCC exams starting 1 PM. Register for exams by September 12. Donation $3.00 adults. Handicap facilities on premises. For information and reservations: Rowena Elrod, KA8OBS, 111 Lancelot Place, Lansing, MI 48906. (517) 482-9650.

NEW JERSEY: October 4. The Orange County ARC will hold its Hamfest and Auction, John S. Burke Catholic High School, 9 AM to 3 PM. Tailgating $3.00. Setup 8 AM. License exams starting 9 AM. Admission $3.00. Tables $7.00. Talk in on 146.76 Rpt. and 146.52 simplex. For more information call Bob, WB2ENA (201) 767-6698.

OPERATING EVENTS

"Things to do . . ."

COLUMBUS DAY SPECIAL EVENT. October 11, 12 1400Z to 2400Z. Open to all Amateur Radio oeprators worldwide. To promote explorer Chkristopher Columbus, the city of Columbus, Ohio and Amateur Radio. Exchange name, QTH and RST. All requests for awards and correspondence shoudl be sent to: ARS W8TO, Special Event Coordinator, 280 East Broad Street, Columbus, OH 43215.

HELP CELEBRATE The Virginia Beach Amateur Radio Club's 25th anniversary by contacting Club Station WA4TGF. Saturday, Sunday, October 5 and 6. For a special 8 x 10 QSL certificate, send 9x12 envelope and two 1st class stamps to R.C. Brown, VBARC, 4821 Rosecroft Street, Virginia Beach, VA 23464.

LONDON BRIDGE Amateur Radio Association, WB7DSW, will operate from Lake Havasu City, the home of the London Bridge, October 11 from 1600Z to 2000Z to celebrate the dedication of the London Bridge. For certificate send QSL to Don Harrington, WB7ALO, 1208 McCulloch Blvd South, Lake Havasu City, AZ 86403.

October 11 and 12: Members of the Dalton ARC, Dalton, GA will operate a special event station from historic Praters Mill Country Fair. The station will be set up in a turn of the century Cotton Gin. 10 to 3 EDT each day. QSL's will include information and a picture of the mill. Dalton ARC, PO Box 143, Dalton, GA 30722.

W.E.C.A. (Westchester Emergency Communications Association) initiates Equipment/"Elmer" Banks to help new hams get on the air. More information write the club at PO Box 131, North Tarrytown, NY 10591.

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Model 610 scope memory

Sibex, Inc. has introduced the Model 610 scope memory, an addition to the line of instruments that convert an analog oscilloscope into a digital storage scope. The Model 610, with its 10-MHz maximum sampling rate, is capable of storing sine wave signals up to 1 MHz in its 2K x 8 static RAM. The input sensitivity is 10 mv/div.

The unit features pre- and post-triggering capabilities or the selection of a 1/2-pre and 1/2-post data combination. A sweep function allows the operator to sweep through the entire memory, then expand any portion of the waveform for detailed analysis.

Priced at $985, Model 610 is available from Sibex, Inc., 3320 U.S. 19 N., Suite 410, Clearwater, Florida 33519

Circle #312 on Reader Service Card.

rechargable memory back-up batteries

Plainview Electronics has introduced its PMB series of rechargeable nickel cadmium batteries for printed circuit board-mounted memory back-up. They're available in 2.4, 3.6, and 4.8-volts, with capacitance of either 35 mAh or 110 mAh. The batteries measure only 16.5 and 23.5 mm, respectively, in diameter.

These batteries are a direct second source to Motorola's MMB series of batteries. Availability is stock to 6 weeks. For information, contact Plainview Electronics, 28 Cain Drive, Plainview, New York 11803.

Circle #309 on Reader Service Card.

mlcroloop™ monoband antennas

Advanced Design Networks (ADN) has announced a new generation of compact monoband antennas called *Microloop*. Across-the-loop dimensions vary from 27 inches for 10 meters to 104 inches for 40 meters. Performance is said to be approximately equivalent to a full-size dipole in a much smaller space.

A comprehensive overview of the *Microloop* is available to readers on request. For information, contact Adavanced Design Networks, 8601 66th Street N., Pinellas Park, Florida 33565.

Circle #308 on Reader Service Card.

THE GUERRI REPORT

by Ernie Guerri, W6MGI

electronic sleuths detect fatal flaws

A high-performance F-15 fighter streaks toward a target, its pilot depending on radar for accurate environmental data. But suddenly the radar picture changes — and then returns to "normal."

Was the pilot in error, or did the radar give erroneous information?

Unfortunately, this was not a simulation, but a real-life situation confronted by radar designers at Hughes. Apparently an intermittent problem in the radar was causing incorrect responses in testing. Worse yet, the problem would also appear in the field, then disappear without explanation.

To solve the problem, Hughes turned to a group of "electronic pathologists" at a division of Gould in Los Angeles. An exhaustive "autopsy" on hundreds of hybrid modules revealed that tin from the plating on the hermetic covers of the modules was contaminating the circuit. The underlying cause was subtle: chemical reactions within the hybrid caused tin "whiskers" — only 0.0001 inch in diameter and about 0.1-inch long — to "grow" inside the module. At random intervals, these whiskers would detach themselves from the case and fall into the active circuitry, causing intermittent operation of just one portion of a complex hybrid. Moreover, random vibration, or current through the whisker, would cause the problem to disappear — frequently without damage to the hybrid.

This was a difficult problem because the whiskers don't begin growing until three to seven years after manufacture, and therefore cannot be detected by any type of screening during the manufacturing process. The only solution may be to design Hi-Res modules with hermetically-sealed packages that don't use plated materials. Unfortunately, this will mean greatly increased complexity and expense.

As the first electronic "MDs" appear on the scene, will it be only a matter of time before the local TV repair shop becomes a "clinic?"

high-power satellites may affect terrestrial systems

There is growing concern that increased transmitter power on some satellites may begin to affect certain terrestrial activities that depend on low power to accomplish their tasks. A particularly important case in point are the various test ranges operated by NASA and the Department of Defense. These ranges typically use signals in the 2.2 GHz region for telemetry links. Missiles and targets are usually very limited in the amount of power they can spare for telemetry transmitters, and the typical transmitter is not more than 3 to 5 watts, with a very simple antenna. The antenna, which usually conforms to the shape of the vehicle's skin, may have little gain (or even a slight loss) with respect to a dipole.

The major concern is that signals from high-power satellites may mask or interfere with these low-power measurement signals, causing erroneous results. Since some tests may require months of planning and cost several million dollars, the consequences of not gathering the intended data are potentially quite serious. It may be possible to coordinate results with the known passing of a given satellite if it belongs to an entity willing to cooperate. However, not all satellite owners and operators have our best interests at heart.

One possible remedy involves using range antennas with more carefully-shaped patterns. Unfortunately, this runs counter to the desire of most range operators, who would like to "look at" the maximum amount of space, to make sure they don't miss anything if a vehicle departs abruptly from its planned course. Another possibility involves establishing polarization standards for satellites and ground stations using the same frequency assignments. While this would add considerable complication to the telemetry ranges, it would be an effective cure — if everyone were to cooperate.

Cooperate? Sounds like the big guys have the same problems in their bands that we see every day on the Amateur bands!

new super-speed semiconductors

Continuing advances in semiconductor technology have been made possible by parallel progress in fabrication techniques. More than 20 years ago it was shown that very high switching speeds could be achieved at room temperatures by an atomic energy-state process known as electron tunneling.

Developed by Esaki in Japan, the theory was brought to practical maturity at IBM and GE in the United States. The early results were the germanium tunnel diodes of the 1960s. In those days, these were the only semiconductor devices capable of oscillating and providing gain in the microwave region.

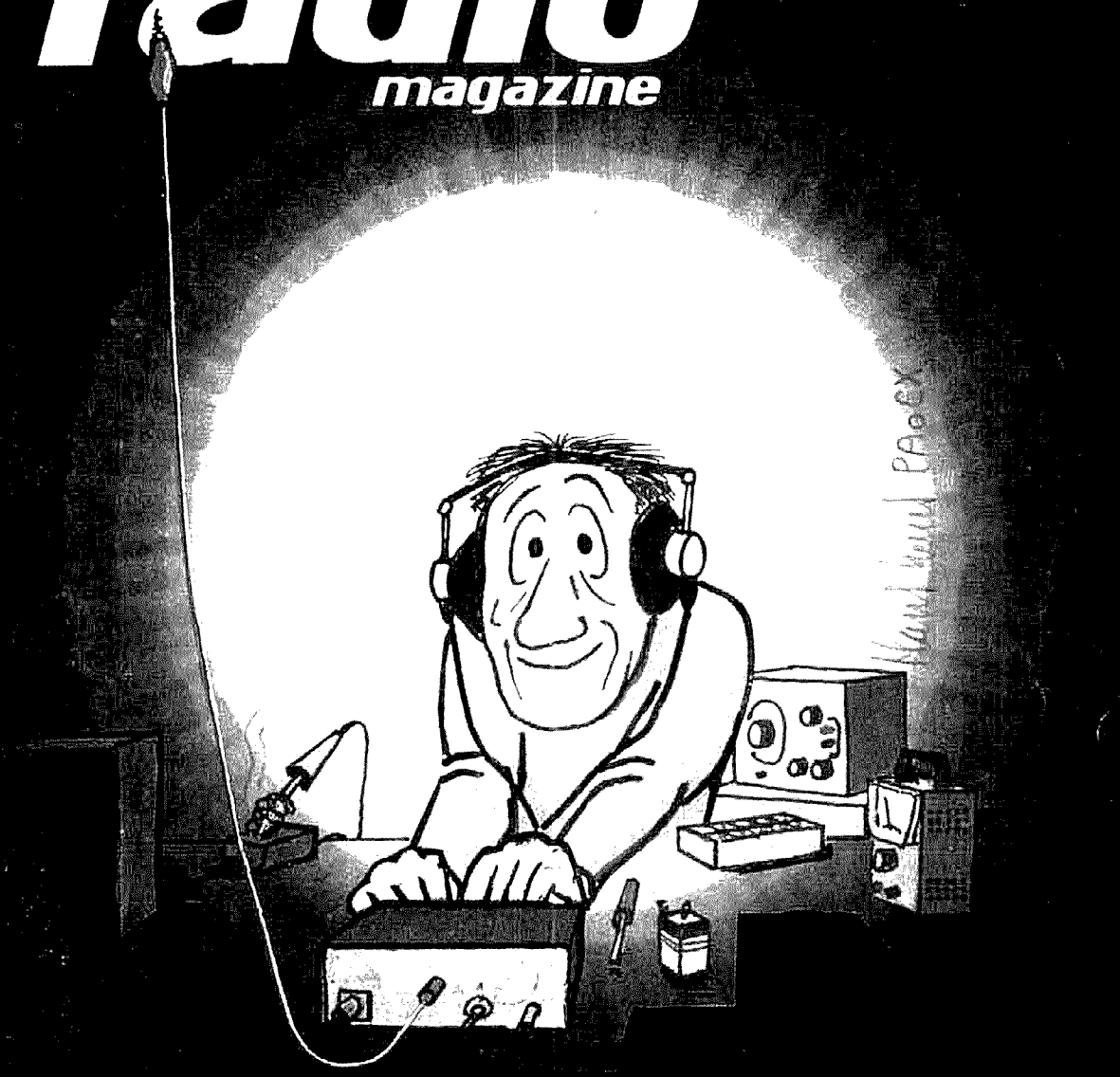
Modern fabrication methods have made it possible to assemble very complex, multi-layer transistors which embody the Esaki concept. The results are still being investigated, but these "resonant tunneling" devices, as they are called, offer the prospect of switching speeds of as fast as 0.1 picosecond (a tenth of a trillionth of a second). Experimenters have reported observing the phenomena at frequencies of about 2.5 THz (1 THz = 1000 GHz).

Don't expect to see commercial devices using these technologies for a long time. Although the ultra-low-noise HEMT has been in the labs for nearly ten years, the first commercial devices were offered on the market just this year.

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Rich Rosen, K2RR
editor-in-chief
and associate publisher

Dorothy Rosa, KA1LBO
assistant editor

Joseph J. Schroeder, W9JUV
Alfred Wilson, W6NIF
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publishing staff

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assistant publisher

Rally Dennis, KA1JWF
director of advertising sales

Dorothy Sargent, KA1ZK
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Susan Shorrock
circulation manager

Therese Bourgault
circulation

cover art:
Hans Evers, PA0CX

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We're sorry to report that pressing international business commitments will make it impossible for W6MGI to continue his monthly column, "The Guerri Report." But we're glad to say that Ernie has promised to take time from his busy schedule to write occasional articles. Watch these pages... Ed.

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REFLECTIONS

examining the EPA's rf radiation limit proposals: what do they mean?

The United States Environmental Protection Agency, concerned with the public well-being, recently released a notice that explores options to prevent adverse health effects that may be associated with exposure to rf radiation. Printed in the July 30, 1986 Federal Register, pages 27318-27339 this notice presented an historical overview of previously proposed and/or accepted radiation limitations (for example, the ANSI standards of 1966 and 1982) as well as reports of various studies involving experiments on laboratory animals. Most importantly, it built a case for a new set of levels to be considered for implementation in this country.

There are important reasons for this action at this time. First, several states have independently enacted legislation setting their own limits on what they consider to be acceptable radiation levels. A uniform nationwide standard, especially one that utilizes the considerable facilities and expertise of the responsible federal agencies, would appear to be a better approach. Second, as a result of growth in the communications industries, public exposure to all forms of rf radiation is greater than ever before. (According to the recent EPA notice, there are approximately 4600 a-m stations, 4400 fm stations and 1100 tv stations in the United States.)

But what is an acceptable level? Conversely, what is too high? The question is simple enough, but the answer depends on many variables. One of the first standards (ANSI, 1966) indicated that a power density of 10 milliwatts per square centimeter across the entire rf spectrum was a reasonable limit. In 1982 ANSI released a revised standard that took into account different levels of absorption versus frequency in the human body, which resonates somewhere between 30 and 300 MHz, depending on size, shape factor, and polarization. The revised standard called for a power density level of 1 milliwatt per square centimeter in that particular frequency range.

As a result of continuing research and the need for a nationally-accepted standard, the EPA is calling for comments on the next generation of proposed acceptable levels. This time, a new term has been defined: SAR or Specific Absorption Rate (watts/kilogram). Mathematically a function of tissue conductivity, electric field strength in tissue, and tissue density, it can be related to our old units of measure — for example, power density and electric and magnetic field intensities.

The EPA proposal discusses four options. The first three indicate specific levels of SAR; the fourth basically sets no level but stresses the importance of continued research. The least stringent option, Number 3, calls for a maximum SAR not to exceed 0.4 watts per kilogram of mass. I believe that's roughly equivalent to our old ANSI standard of 10 mw/cm^2. Options Number 2 and 1 are correspondingly more stringent, calling for SARs not to exceed 0.08 and 0.04 watts/kg, respectively.

There is considerable evidence to indicate that above an SAR of 4 watts/kg, the core temperature of the human body increases, leading to health complications. Notice that the EPA's option Number 3 (the least stringent) is one tenth of this amount. However, pre-existing health conditions such as poor circulation and normally elevated body temperatures reduce the level of acceptable SAR.

At this point you're probably wondering, "How does this affect me?" It's a good question, but I'm not sure of the answer. The EPA specifically proposed (on page 27334) to exclude radiation from "consumer electronic products" from the proposed standard. I think that category includes our transmitters, linears, and other equipment. The EPA's primary aim is to protect the public from radiation from high-power rf sources — mainly a-m, fm, and tv stations running many kilowatts. But perhaps this might be added incentive for the few "superpower boys" out there to either cut back or move their families away from the high rf fields.

In this short space it's not possible to cover in detail many of the points brought out in the EPA notice. I highly recommend that interested readers obtain a copy of these 22 pages and specifically review several of the 62 cited references, as well as an excellent article on the subject ("The Microwave Problem," by Kenneth R. Foster and Arthur W. Guy) that appeared in the September, 1986 issue of *Scientific American*.

The conversions from power source to rf radiation levels to SARs are derivable or well-documented. Through its investigations and those of other agencies, the EPA hopes to bridge that last gap in understanding — the one that would relate SARs to health — and thereby provide a nationwide rf radiation standard that we can live with.

— Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

THE "ELECTRONIC COMMUNICATIONS PRIVACY ACT" WAS MOVING RAPIDLY toward Senate approval at presstime and may even have become law by the time this sees print. On September 19 the Senate Judiciary Committee unanimously approved the latest version of S-2575 after only 25 seconds consideration! Fortunately the Senate version, though still objectionable for its basic philosophy that U.S. citizens can't listen to any service or frequency that the government doesn't want them to, has been even more tempered than the version passed by the House in June. The bill now permits monitoring any radio communication "for the use of the general public" as well as cordless phones, marine or aircraft, Amateur, CB, GMRS, government, law enforcement, Civil Defense, private land mobile, or public safety (police and fire) communications which are "readily accessible to the public."

Two Last-Minute Amendments Further Limit The Senate Bill by eliminating criminal penalties for intercepting (without malicious intent) broadcast remote pickup stations and by reducing the penalty for first interception of public land mobile services (cellular, older car phones, and voice paging) to a $500 fine. These worthwhile amendments, proposed by the Association of North American Radio Clubs, were introduced by Senator Paul Simon (D--IL).

If All Goes Smoothly The Bill Could Move Quickly through the Senate Subcommittee on Communications and on to the Senate floor, where it could be voted on before the end of the month. However, columnist Jack Anderson's September 23 syndicated column took a very hard swing at the bill, blasting the cellular telephone industry for having sold both Congress and cellular users a bill of goods on the legislation and "cellular privacy." Its very timely appearance could well induce some Senators to raise enough questions about the bill to slow up, if not derail, its passage. Furthermore, though Senator Goldwater hasn't made his views on the bill public, it's considered unlikely to win his endorsement.

A PRIVATELY ADMINISTERED AMATEUR CALLSIGN PROGRAM is looking more and more likely, with not only ARRL but several other groups interested in taking on the job. The idea has actually been under consideration at the Commission for some time, and FCC officials discussed it at length during the FCC Forum at the recent ARRL National Convention.

Under Its Plan FCC Would Issue Only Basic 2x3 Initial Callsigns to newly licensed Amateurs. An Amateur who upgraded could continue to use that callsign or go to the "Callsign Administrator" who -- for a fee -- would issue a new callsign appropriate to the new license class. The FCC estimates charges for each callsign upgrade would be in the $20-30 range. There would also be an initial "grace period" during which those who wanted to recover a previously held callsign or one formerly held by a deceased family member could request and receive that callsign.

The FCC Is Quite Interested In Moving Forward On The Proposal and could decide to solicit comments on it in a Notice of Inquiry or even "PRB-3" before the year ends.

GEO-SYNCHRONOUS "PHASE 4" AMATEUR SATELLITES are being seriously considered by AMSAT. A just-completed "Phase 4 Technical Study Plan" by AMSAT Engineering VP W3GEY proposes a pair of such birds, positioned above the equator so as to provide a "footprint" reaching from Japan and Australia eastward to most of Europe, Africa and the Middle East.

2-Meter Plus 70, 24 And 13 cm Transponders with a wide variety of capabilities would be carried by the birds, which would be aimed at a much broader segment of Amateur Radio than ever before with a strong "Public Service" orientation. It's possible that even mobile or hand-held radios would be capable of working through the new satellites.

Next Step For The Ambitious Proposal Is A Study And Planning Period of a year or so; launch would be targeted for 1990-91. Preliminary cost estimates are in the million-dollar range, meaning that very broad-based international Amateur support will be required.

NASA's 1995 "Space Station" Plan May Include A Built-In Amateur Station as a result of a recent meeting of NASA, AMSAT, and ARRL representatives. Possibilities look very good, and a working group has been set up to make a formal proposal to NASA next year.

OSCAR 10's Salvage Efforts Continue To Encourage Recovery Workers, though an upcoming solar panel eclipse and resulting battery discharge will soon put it temporarily out of service anyway. When sunlight does return to recharge its batteries, the chances that OSCAR 10 will be able to be brought back to useful life are estimated at no better than 50-50.

AMSAT's Fourth Annual Space Symposium And Annual Meeting will be held at the Dallas/Ft. Worth Hilton November 7-9. Call AMSAT at (301) 589-6062 for details.

FIRST RECIPIENT OF THE "AEA AMATEUR AMBASSADOR AWARD" is Mary Duffield, WA6KFA, who was recognized for her work with the Redwood Youth Foundation, which has introduced hundreds of youngsters to Amateur Radio. The retired telecommunications educator received the honor and an accompanying $1000 cash award at ARRL's San Diego National Convention. Congratulations!

U.S. AMATEUR POPULATION IS STILL GROWING, ALTHOUGH VERY SLOWLY. Most encouraging is the dropout rate, which has decreased steadily from an average of 1478 per month in 1984 to 1210 per month in 1985 and only 910 per month so far this year. At the end of August the total U.S. Amateur population was 421,077.



comments

something for everyone

Dear HR:

Credit should be given when it is due. The September issue of *ham radio* is outstanding for my tastes in technical literature. The spread of analog subjects of immediate interest to the dedicated Amateur was just right! Certainly, there should be something for everyone in the September issue.

I was especially attracted to and enlightened by the K1GQ article ("Analyzing 80-meter Delta Loop Arrays," page 10). Having been a loop enthusiast for many years, I found the Myers treatment of the subject to be the most definitive one I have read. His findings confirm many of my intuitive views about the various loop configurations. I am presently using a full-wave rectangular loop for 80 meters, corner-fed and erected vertically. It seems to perform better than the tilted delta loop I used for two years.

The W4UCH article, "Low-cost Spectrum Analyzer with Kilobuck Features" (page 82), was similarly stimulating. I find myself beset with anxiety to build a copy of his unit for use in my lab. The circuit is quite "sanitary" and to the point!

I enjoyed also the neat article by Mike Gruchalla concerning the NE5205 wideband amplifier (page 30). If time permitted such luxury, I would be immersed in component parts and flux fumes for some time, as I built each of the projects that caught my fancy in September *ham radio*. Keep it up, and try to maintain the good mix of topics for those of us who aren't driven by the present-day digital philosophy. The Amateur journals and *rf design* may represent the salvation of the rf/analog technology. The trade journals have surely lost sight of the importance of rf technology!

Doug DeMaw, W1FB/8
Luther, Michigan 49656

safety first

Dear HR:

N1ACH's editorial emphasis on safety is well placed ("Switch to Safety," September, page 4). Every consideration possible should not be overlooked.

As a teenager I experienced the full impact of a misfortune related to this problem. My father was working on a sub-station transformer when a careless operator turned on the main circuit breaker; 13,500 volts of electricity went through his body, resulting in injuries which led to his death. The power company then instituted a procedure of grounding all lines when men were working on them.

Years later, while in the Navy, I almost joined my ancestors by a similar route. I was working on the SJ submarine radar. I had carefully opened the circuit breaker and tagged it with a regulation red warning tag. I also checked for power at the input terminals to the radar. A few minutes later there was a flash of blinding fire and the jaws of my pliers melted. A young, inexperienced officer, in attempting to restore lighting to his room, had closed the circuit breaker. Fortunately, I was wearing thick-soled shoes and never provided a circuit through my body — the old "hand in the pocket" technique.

At our high-power Navy radio stations, we had key switches. When a technician opened the circuit breaker, he locked the switch and took the key with him.

Vigilance is the word. And *never* trust the other guy.

I. L. McNally, K6WX
Sun City, California 92381

price drops on "neat little book"

Dear HR:

As I read my September, 1986, issue of *ham radio* I was pleased to see Cushman Electronics mentioned in "Low-Cost Spectrum Analyzer with Kilobuck Features" by Robert Richardson, W4UCH. Please compliment Mr. Richardson on a very well written technical article.

Incidentally, the "neat little book entitled *Using the Spectrum Monitor*, priced at $7.25 (postpaid)," is now a neat little book priced at $4.95 (postpaid).

Also, our new address is 1525 Atteberry Lane, San Jose, CA 95131.

Ray V. White, KI6DW
President, Cushman Electronics

update on ACSSB level-one adapter

Dear HR:

There have been several small value changes to the Project OSCAR level-one adapter described in the October issue of *ham radio*. We will gladly provide the revisions and pc board information to anyone sending an SASE to Level-One Adapter, 15 Valdez Lane, Watsonville, California 95076.

James Eagleson, WB6JNN
President, Project OSCAR

short circuit

program correction

In October's "DX Forecaster," the exponentiation symbol was omitted from lines 100, 200, 280, and 360 of the information provided for adapting the HORANT program for use on the Apple IIe and IIc. The following lines should be substituted for existing lines in the HORANT program shown on page 92:

```
 95  IF 4*H1 < L1 THEN 420
 98  Y = (L1/(4*H1))
100  A1 = ATN (Y/(1-Y^2)^0.5)
198  Y = (3*L1)/(4*H1)
200  A2 = ATN (Y/(1-Y^2)^0.5)
278  Y = (5*L1)/(4*H1)
280  A3 = ATN (Y/(1-Y^2)^0.5)
358  Y = (7*L1)/(4*H1)
360  A4 = ATN (Y/(1-Y^2)^0.5)
```

ham radio

In this edited excerpt from his forthcoming book, Ulrich Rohde, well known to readers for his many articles on receivers and other communication subjects, shares knowledge borrowed from commercial applications to improve noise handling capabilities in Amateur equipment.*

understanding and handling noise

Balancers, limiters, and squelch circuits enhance receiver performance

Pulse interference is often the limiting noise source for communication receivers on our HF and VHF bands. To understand how pulse interference affects performance — and before noise-reducing methods can be developed — it's necessary to understand the sources and types of interfering pulses. Although many solutions have been proposed since the early days of radio[1], a large part of this information is based on recent work.[2,3,4,5]

types of pulse interference

Noise can be viewed in both the time domain, with an oscilloscope, and in the frequency domain, using a spectrum analyzer.

The clicks that appear when a current is switched on or off represent the simplest form of an interference pulse (**fig. 1A**). In this case the energy in the interference spectrum is a function of the amplitude, slew rate, rise time t_r (i.e., the time it takes to build from 10 to 90 percent of final amplitude) and the repetition frequency. For wideband systems with a low-pass characteristic, the approximate formula $t_r = 0.3/\Delta f$ is valid, where Δf is the bandwidth in Hz and t_r is the rise time in seconds.

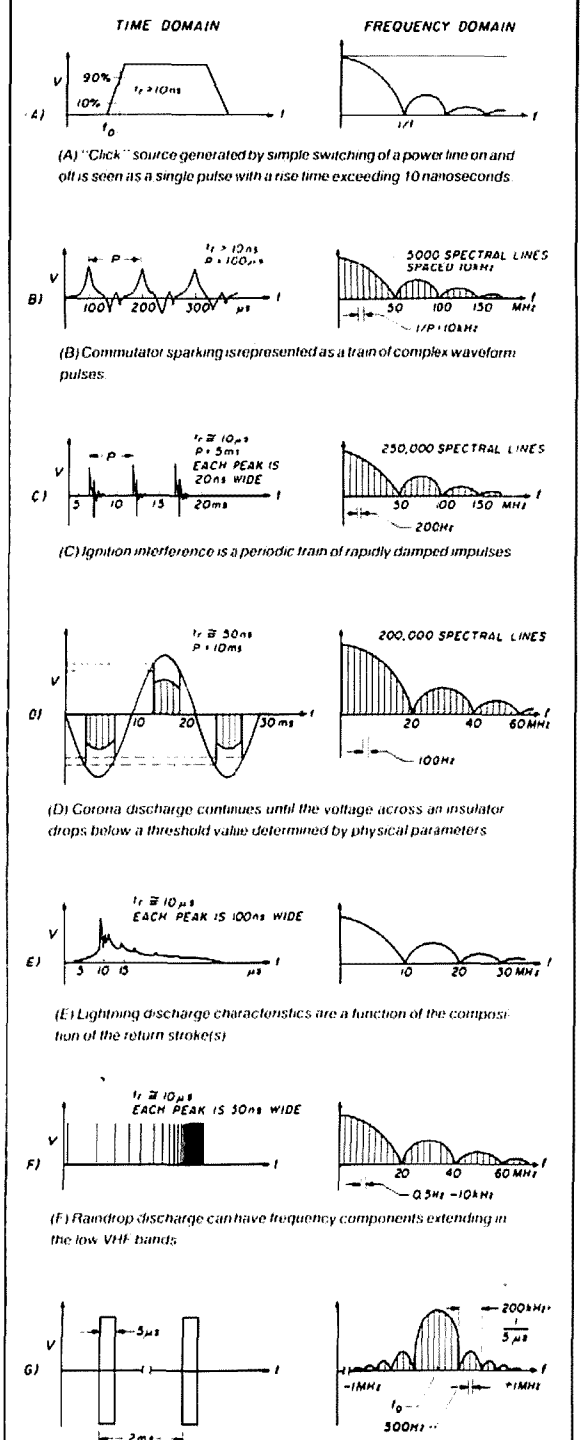
The theoretical vertical step with $t_r = 0$ (infinite rate of rise) has a spectrum that extends to infinity — i.e., the "click" would be heard at all frequencies. For finite pulse duration, t, the amplitude characteristic of the spectrum has a shape described as the sine integral (**fig. 1A**). As t increases, the frequency representation of the noise compresses — i.e., the minima in the spectrum are shifted toward lower frequencies. In actual circuits t_r will be at least 10 nanoseconds because of the presence of some line inductance. This results in the main spectrum components of this form of interference being below 30 MHz. Amplitudes are low and decrease above this frequency.

Commutator sparking is a special case of switching interference, in that a series of individual switching pulses are generated according to the type of machine, the number of poles, and speed (**fig. 1B**). Repetition frequencies up to several kHz can occur. This results in a higher spectral energy density than in the case of the single switch, proportional to this frequency factor. Thus, sometimes the rotating machine noise can be heard up to 30 MHz. The spectral amplitudes can increase by several orders of magnitude when arching occurs between the commutator and brushes. However, this does not occur often, since it would quickly destroy the commutator.

Ignition interference, which results from a high voltage discharge in a pressurized gas, has a repetition frequency of less than 1 kHz and a spectrum as indicated in **fig. 1C.** In the case of conventional circuit-breaker type ignition systems, the voltage amplitude is highest and the rise time shortest at low engine speeds. The breaker points, together with the high-voltage cables connected between the spark plugs and distributor, form resonant circuits which act like the

**HF Communications Receivers: Theory and Design*, by U.L. Rohde, Ph.D., and T.T. Nelson Bucher, Ph.D., to be published by McGraw-Hill Book Company, New York, 1987. Used with permission.

By Ulrich Rohde, DJ2LR, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458



(A) "Click" source generated by simple switching of a power line on and off is seen as a single pulse with a rise time exceeding 10 nanoseconds.

(B) Commutator sparking is represented as a train of complex waveform pulses.

(C) Ignition interference is a periodic train of rapidly damped impulses.

(D) Corona discharge continues until the voltage across an insulator drops below a threshold value determined by physical parameters.

(E) Lightning discharge characteristics are a function of the composition of the return stroke(s).

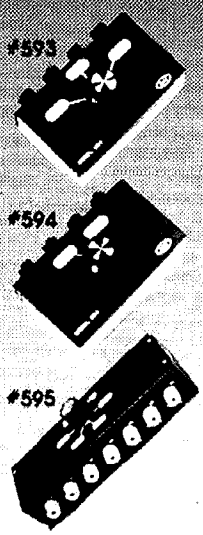
(F) Raindrop discharge can have frequency components extending in the low VHF bands.

(G) Radar pulses have rise times that can vary from 10 nanoseconds to 1 microsecond.

fig. 1. Time and frequency domain representations of seven different types of "noise" interference. (Courtesy of M. Martin and *VHF Communications*, Spring, 1980.

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spark transmitters used in the early days of radio communications. The oscillations decrease in an exponential manner.

Corona discharge occurs across defective insulators on high-voltage ac lines. A discharge can be ignited which, depending on the "quality" of the insulator, starts more or less below peak voltage and lasts until the voltage is reduced below a threshold which is always lower than the starting voltage. This process is repeated during the opposite half waves of the alternating voltage. (See **fig. 1D.**) In the case of a poor three-phase line, or even with a good line in very high humidity (for example, in fog), arcing of all three phases could result in an overlap of the burning times (which occur every 3.33 milliseconds for a 50-Hz line and every 2.78 milliseconds for a 60-Hz line.[6]

Lightning discharges account for what is called "atmospheric noise." In relatively flat terrain, 85 percent of all lightning strokes occur in a downward direction. The cloud involved is negatively charged and a current surge of up to 20 kA passes between it and the earth. Eighty percent of these lightning flashes have maximum currents of 50 kA.[7] It is, however, possible for several clouds to discharge via the same ionized path, in which case currents can exceed more than 200 kA. (See **fig. 1E.**) The duration of a lightning discharge can be between 0.1 milliseconds and 0.5 second, depending on how many subsequent surges are induced by the primary discharge.

Raindrop discharge, also known as precipitation static, results from charges transferred between a grounded antenna and statically charged raindrops. It causes a click-type interference which can be heard as an occasional individual click or a uniform white noise during heavy showers. Because of the very steep slope of this discharge, very high interference amplitudes are generated in the resonant ranges of the antenna. Although the capacitance of a single raindrop is only a few pF and is rapidly discharged, the high potential and large repetition frequency produce a considerable energy density, as indicated in **fig. 1F.** The effect is especially prevalent on the antennas of aircraft flying through cloud formations. A similar effect can be produced by sand or dust storms.

Pulses from radar transmitters that operate in the GHz range (as well as lower frequency over-the-horizon radars) are another source of noise. These transmitters use relatively short pulses (**fig. 1G**) and, depending on the shape of the keying waveform modulation, generate sidebands with various frequency (amplitude) spectra. Similar to the case of baseband pulses, the spectral line spacing of rf pulses is determined by the repetition frequency. Zero spectrum points occur at the reciprocal of the pulse length and its multiples.

effects of pulses on narrow-band systems

A narrow-band system with tuned (resonant circuit) amplifiers only enhances frequency components that fall within its passband. An individual resonant circuit is "shock"-excited to oscillation at its resonant frequency by an impulse slope. The transient duration is dependent on the bandwidth, as determined by the circuit Q. In the case of multistage amplifiers or multipole filters, the output pulse delay is dependent on the circuit group delay, t_g. This increases linearly with the number of resonators and is generally measured from the input impulse time, t_o to the time at which the output signal has risen to 50 percent of its peak amplitude. An approximate formula gives $t_g = 0.35N/\Delta f$, where N is the number of resonant circuits, Δf is the bandwidth (Hz) and $t_r = 1/\Delta f$. If different bandwidth stages are cascaded, rise time is determined mainly by the narrowest filter. The group delay results from the

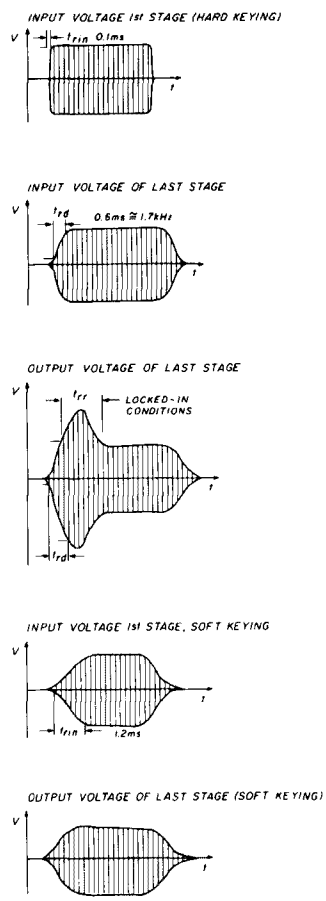


fig. 2. Impulse drive of an rf amplifier with AGC. (Courtesy of M. Martin and *VHF Communications*, Spring, 1980.)

sum of the individual delay times.

If short rise time rf impulses are fed to an amplifier that has a much longer rise time, three different types of responses can occur.[8]

- If the input pulse has duration, t_p, longer than the transient time t_{rv}, the output signal achieves the full amplitude and maintains it for the duration of the drive time less the transient time $t_p - t_{rv}$.
- If t_p is equal to t_{rv}, the output signal achieves full amplitude during time t_{rv}, but then immediately starts to decay to zero in a period, t_{rv}.
- If the pulse is shorter than the circuit transient time, the response is essentially that of the transient, but with an amplitude which only reaches a portion of the pulse; i.e., the shorter the pulse, the smaller the amplitude produced by the amplifier. In other words, a substantial portion of the spectrum of the input pulse is not within the bandwidth of the amplifier, and, therefore, does not contribute to the output amplitude.

In many systems an AGC circuit is provided to ensure that a large range of input signal voltages is brought to the same level at the demodulator output. The AGC control voltage is generated subsequent to the narrow-band i-f filter. This filter might have a bandwidth of from approximately 2 to 12 kHz, corresponding to transient times of about 0.2 to 1 millisecond. Earlier stages have much broader bandwidths. This means that steep input pulses can drive earlier amplifier stages into saturation before a reduction in gain is effected by AGC (whose response time is generally longer than the narrow band filter transient response). This is especially true in superheterodyne receivers, where selectivity is determined in the final i-f, often after substantial amplification of the input signal. Thus, the audible interference amplitudes may be several times stronger than the required signal level after the AGC becomes effective. A long AGC time constant worsens the situation. It is only when the rise time of the input pulse is longer than the AGC response time (e.g., during telegraphy with "soft" keying) that the output amplitude will not overshoot, as shown in **fig. 2.**

An effective method of limiting the maximum demodulator drive, and thus reducing the output pulse peak, is to clip the signal just in front of the demodulator with symmetrically limiting diodes, so that the i-f driver amplifier is not able to provide more than the limited output level — e.g., 0.7 volts peak-to-peak. It is necessary in this case that the AGC detector diode be delayed no more than a fraction of this level (corresponding to 0.4 volts) so that the clipping process does not interfere with AGC voltage generation. Use of a separate AGC amplifier, not affected by the limiter, will also assure this, and can provide an amplified AGC system to produce a flatter AGC curve.

methods for suppressing interfering pulses

Three different methods are used to suppress interfering pulses:[9]

- *Balancers* attempt to reproduce the pulse shape without the signal in a separate channel, and then perform a subtraction from the channel containing both signal and pulse.
- *Limiters* attempt to prevent the pulse level from becoming excessive.
- *Blankers, or silencers,* attempt to detect the onset of a pulse, and reduce to zero the gain of the signal amplifier chain, at an early stage, for the duration of the pulse.

Balancer systems are designed to obtain two signals in which the signal and noise components bear a relatively different relationship to each other. The two signals are connected in opposition so as to eliminate the noise while keeping the signal. The main problems with this type of impulse noise suppression are obtaining suitable channels and exactly balancing out the noise impulse, which is generally many times stronger than the desired signal.

In commercial service, attempts have been made to use different frequency channels with identical bandwidths to get the same impulse shape, but with the signal in only one channel. The difficulties of matching channels and finding interference-free channels makes this approach unsatisfactory in most cases. Other approaches have attempted to slice the center from the pulse (to eliminate the signal) or to use a high-pass filter to pass only the higher frequency components of the pulse for cancellation. While some degree of success can be achieved with such circuits, they generally require very careful balancing and hence are not useful when a variety of impulses and circuit instabilities are encountered.

This type of circuit can be useful where the impulse source is a local one which is physically unchanging. In this case, a separate channel from the normal antenna can be used to pick up the pulse source with negligible signal component, and the gain of the pulse channel can be balanced carefully using stable circuits (and a feedback gain control channel if necessary). It has also been reported that modern adaptive antenna systems with sufficiently short response times can substantially reduce impulse noise coming from directions other than the signal direction.

Since most of the noise terms contain relatively large peaks, limiters have been used, especially in a-m reception (which includes, of course, telegraph through full-carrier voice signals) to clip audio signal peaks which exceed a preset level. **Figure 3** shows a series limiter circuit at the output of an envelope demodulator which has proven effective in reducing

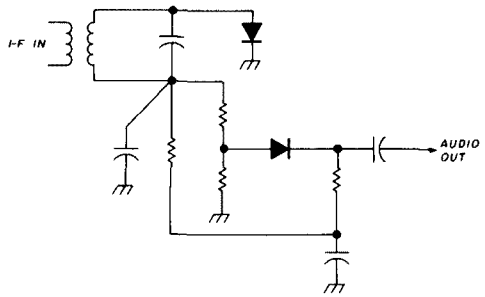


fig. 3. Schematic diagram of automatic series noise limiter circuit.

audio noise caused by impulse interference. This type of circuit makes listening to the signal less tiring, but does not improve intelligibility of the received signal. The limiting level may be set to a selected percentage of modulation by adjusting the tap position of the two resistors feeding the limiter diode's collector. If set below 100 percent, the limiting level also limits peaks of modulation. Since these seldom occur, such a setting is acceptable.

Because the impulse amplitude is higher and the duration shorter in the early stages of a receiver, limiting in such stages reduces the noise energy with less effect upon signal than limiting in later stages. Some fm receivers use i-f stages which are designed to limit individually, while gradually reducing bandwidth by cascading resonant circuits. Such a design eliminates strong short impulses early in the receiver, before they have had a chance to be broadened by the later circuits. Such receivers perform better under impulse noise conditions than those which introduce a multipole filter early in the amplifier chain. However, wideband limiting can reduce performance in the presence of strong adjacent channel signals.

The principles discussed are also applicable to data receivers. While the impulse interference is stronger than the signal, the signal modulation contributes little to the output. Generally, the data symbol duration is longer than the duration of the input impulses. If the impulse can be reduced or eliminated before the establishment of final selectivity, only a small portion of the signal interval is distorted, and a correct decision is much more likely. Consequently, limiters at wide bandwidth locations in the amplifying chain can result in considerable reduction of error rate in a data channel. Again, the possibility of interference from adjacent or other nearby channel receivers must be considered.

Impulse noise blankers are based on the principle of "opposite modulation." In effect, a stage in the signal path is modulated so that the signal path is blanked by an amplitude modulation process for the duration of the interference. It is also possible to use a frequency modulation method in which the signal path is shifted to a different frequency range. This latter procedure[10] uses the attenuation overlap response of i-f filters in a double superheterodyne. The second oscillator is swept several kHz from nominal frequency for the duration of the interference so that the gain is reduced to the value of the ultimate selectivity in accordance with the slope of the filter curves. This method is especially useful since the switching spikes which often accompany on-off modulation should not be noticeable. However, when using an fm modulator having high speed (wide bandwidth), components can appear within the second i-f bandwidth from the modulation. The most stringent limitation of this method is the requirement for two identical narrowband filters at different frequencies along with an intermediate mixer. This technique is limited to a double conversion superheterodyne with variable frequency first (local) oscillator.

When using an amplitude modulation method, two types of processing are possible:

(A)The interference signal is tapped off in parallel at the input of the system and increased to the trigger level of a blanker by an interference channel amplifier having a pass bandwidth which is far different

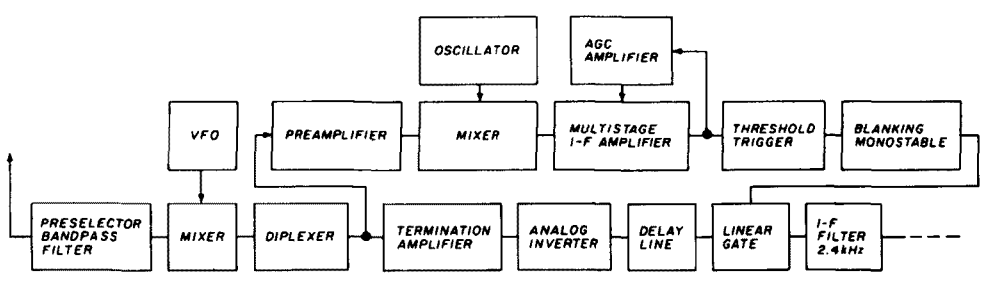


fig. 4. Block diagram of superheterodyne with noise blanker.

from the signal path. This method is effective only against very wideband interference, since noticeable interference energy components must fall into the passband range to cause triggering. This method will not be effective in the case of narrow-band interference, such as radar pulses, which are within or directly adjacent to the frequency range to be received.

(B)The interfering signal is tapped off from the required signal channel from the mixer and fed to a fixed-frequency, second i-f amplifier, where it is amplified up to the triggering level.[2,3] Since there is danger of crosstalk from the interference channel to the signal amplifier channel, it is advisable to have a frequency converter stage in the interference channel. This would process the interference at a different frequency than the signal i-f. In using this method, make sure that no switching spikes generated during the blanking process are fed back to the interference channel tap-off point. If they are, there will be danger of pulse feedback. The return attenuation must, therefore, exceed the gain in the interference channel between the tapping point and the blanker.

The blanker must be placed ahead of the narrowest i-f filter in the signal path. It must provide blanking action before the larger components of the transient have passed this filter. Therefore, we must assure a small group delay in the interference channel by use of a sufficiently broad bandwidth and a minimum of resonant circuits. It is desirable to insert a delay between the tap-off point and the signal path blanker so that there is sufficient time for processing the interference signal. If this is done, it is not necessary to make the interference channel excessively wide, while still assuring the suppression of the residual peak.

Figure 4 is the block diagram of a superheterodyne that uses this type of impulse noise blanker. **Figure 5** illustrates its operation in the presence of a strong interfering radar pulse. An essential part of the blanker is the use of a gate circuit that can operate linearly over a wide dynamic range. **Figure 6** shows such a gate, using multiple diodes. The circuit is driven by a monostable flip-flop which is triggered by the noise channel. **Figure 7** is a schematic of a noise blanker circuit designed along these lines. When the noise channel is wider than the signal channel, this type of noise reducer can also have problems from interfering signals in the adjacent channels.

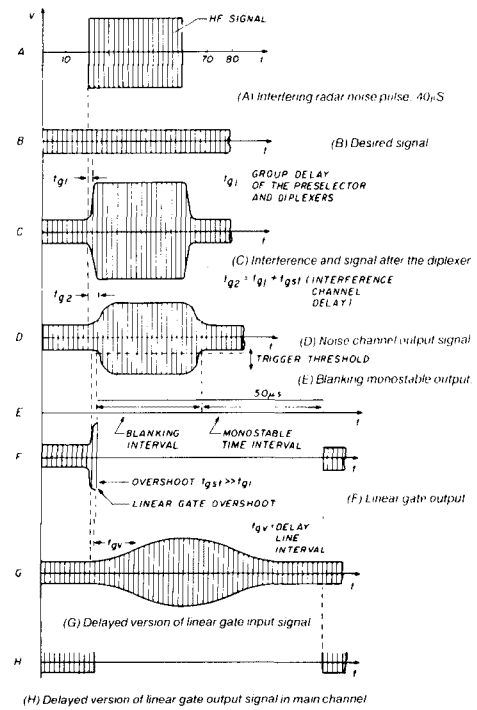


fig. 5. Observed signal waveforms in a superheterodyne receiver that uses an impulse noise blanker.

squelch circuits

Sensitive receivers produce considerable noise voltage output in the absence of a signal. This condition can occur when tuning between channels or when the station being monitored has intermittent transmissions. At the audio output the noise can be annoying and, if repeated frequently, fatiguing. To reduce this problem, circuits are often provided to reduce the output when a signal is not present. These circuits have been referred to as *squelch*, *muting*, and *quiet AVC (QAVC) systems*. The choice of which circuit to use depends on the received signal characteristics.

Squelch circuits for a-m receivers generally operate from the AGC control voltage. When a weak signal or no signal is present, the voltage on the AGC line is at its minimum and receiver gain is maximum. When a usable signal is present, the AGC control voltage rises to reduce the receiver gain. The voltage variation tends to rise approximately logarithmically with increasing signal levels. By using a preset signal level threshold, it is possible to gate off the audio output signal whenever the signal level drops below this point. Such a system can be used to mute the receiver during the tuning process. The threshold may also be set for the level of a particular signal with intermittent transmissions, so that noise or weaker interfering signals are not heard when the desired signal is off. When the transmission medium causes signal fading, as is common at hf, squelch circuits are somewhat less effective for this use because the threshold must be set low enough to avoid squelching the desired signal during its fades. This provides a smaller margin to protect against noise or weaker interfering signals.

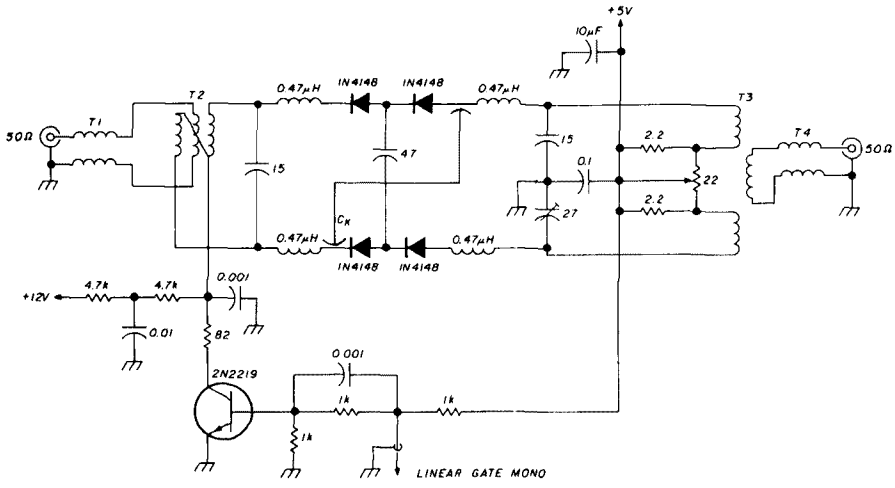


fig. 6. Schematic diagram of blanker gate with high dynamic range.

A typical a-m squelch system is illustrated in **figs. 8** and **9**, where a diode gate is used to reduce the output signal. Many types of switches have been used for this purpose, including biasing the demodulator diode and biasing one element of a multi-element amplifying device. The latter approach was frequently used with receivers that employed multigrid vacuum tube amplifiers. However, it can be applied to multigate FET amplifiers or balanced amplifier ICs with current supplied by a transistor connected to the common base circuit of the amplifying transistor pair. **Figure 10** shows these alternate gating techniques.

Many fm receivers do not use AGC circuits, but depend on circuit limiting to maintain the output level constant from the demodulator. In this case, squelch may be controlled by the variations in voltage or current which occur in the limiter circuits. Such changes occur when single-ended amplifiers are used for limiting, but may not be so readily available in balanced limiter arrangements. Furthermore, the wide range of threshold control provided by AGC systems is generally not available in limiters. This tends to make fm squelch systems, which are dependent on the signal level, more susceptible to aging and power supply instabilities than AGC operated systems. Consequently, two other types of control have evolved for fm use: noise-operated and tone-operated. (The latter could be used for a-m, also.)

Figure 11 is a block diagram of a noise-operated squelch. This system makes use of the fact that output noise characteristics from a frequency demodulator change when no signal is present. At the low output frequencies, when noise alone is present in the fm demodulator, there is a high noise level output, comparable to that at other frequencies in the audio band. As the strength of the (unmodulated) signal rises, the noise at low frequencies decreases, while the noise at higher frequencies decreases much less rapidly.

If in **fig. 11**, the squelch low-pass filter cuts off at, say, 150 Hz, it will be uninfluenced by modulation components. If the gain of the squelch amplifier is set so that the squelch rectifier produces 5 volts when S/N = 0, then a 7-dB signal level will cause this output to drop to about 0.03 volts. A threshold may readily be set to cause the squelch gate to open at any S/N level between −3 dB and 7 dB. Because the control voltage level is dependent on the gain of the rf, i-f, and squelch amplifiers, variations in squelch threshold may occur as a result of gain variation with tuning or because of gain instabilities. If a second filter channel (**fig. 12**) tuned above the baseband is used, the two voltages can be compared to key the gate on. While both are subject to gain variations, their ratio is not. Better threshold stability results. A similar technique has been proposed for SSB voice, where noise density is uniform, but modulation energy is greater below 1 kHz.

With the difference approach, the range of threshold control is, however, limited. A weak interfering signal which would produce negligible interference with the desired signal may still operate the squelch gate. Tone-operated squelch was devised to overcome this problem. A small-deviation, low-frequency tone (lower than the modulation frequencies) is added to the transmitted signal. At the receiver, a narrow-band filter is tuned to the tone, and its output is amplified and rectified to operate a trigger for the squelch gate.

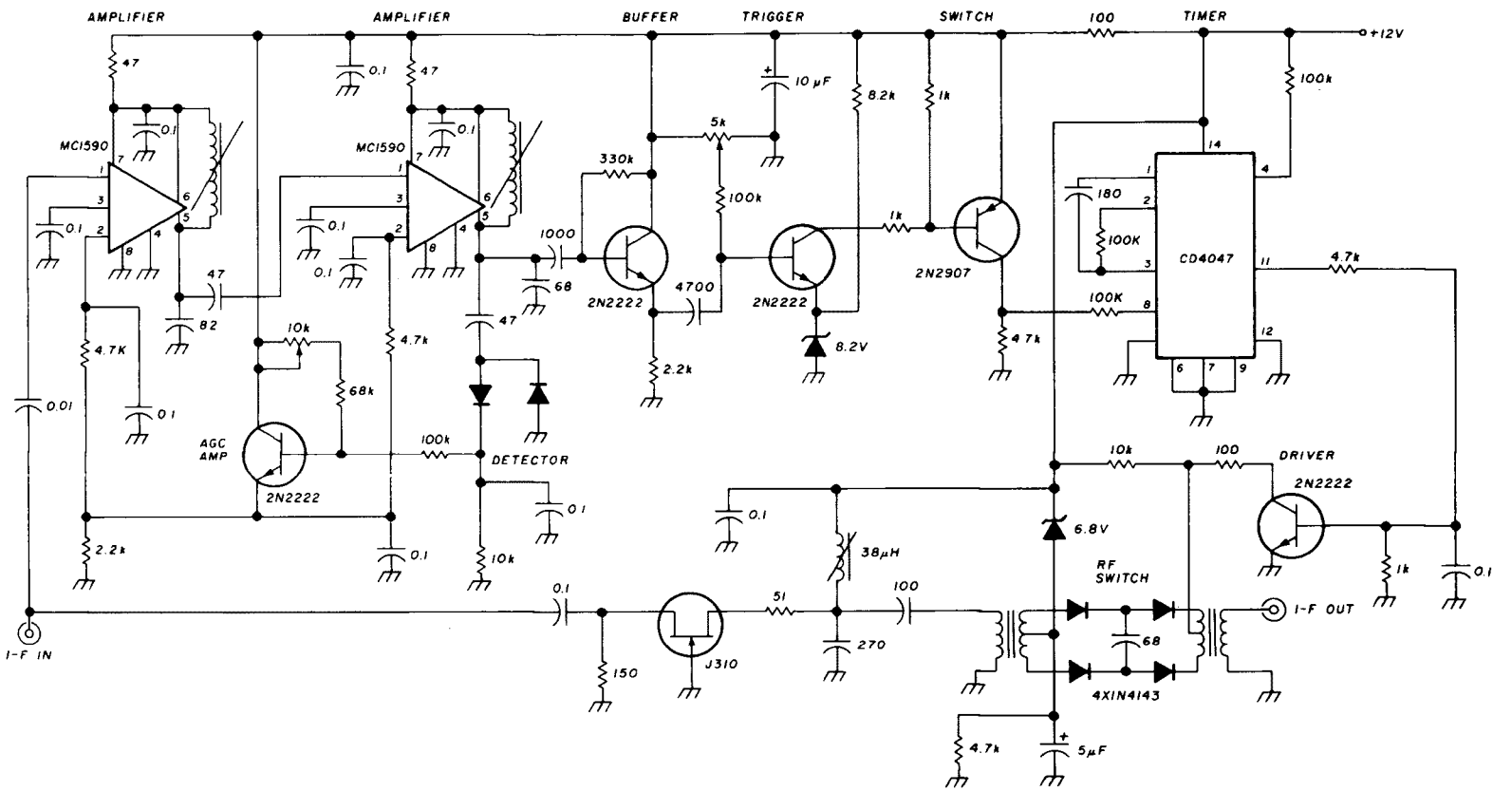


fig. 7. Schematic diagram of noise blanker circuit.

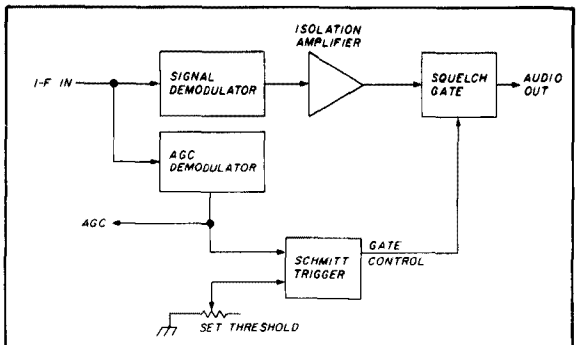


fig. 8. Block diagram of a-m squelch circuit.

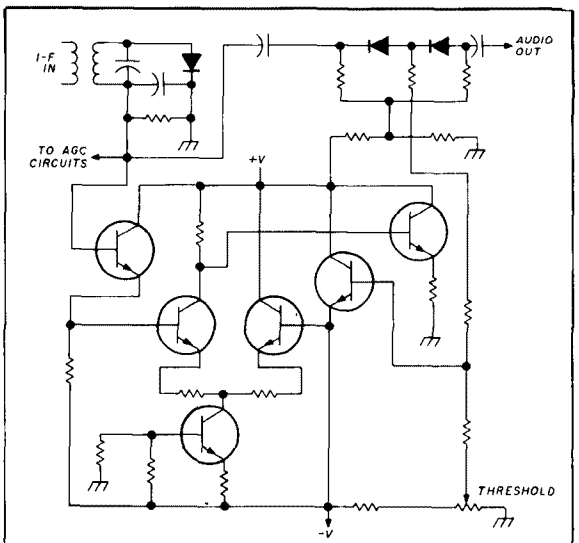


fig. 9. Schematic diagram of a-m squelch circuit.

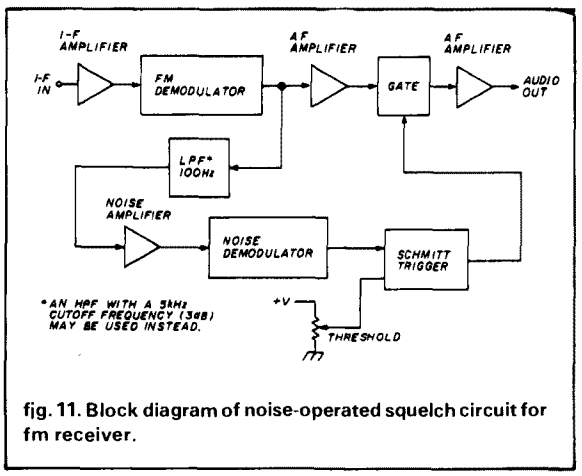


fig. 11. Block diagram of noise-operated squelch circuit for fm receiver.

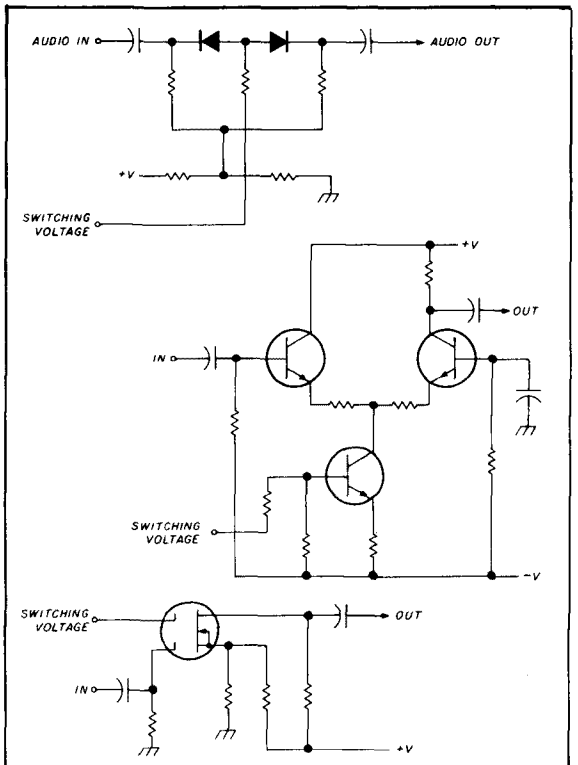


fig. 10. Some gate circuits for squelch application.

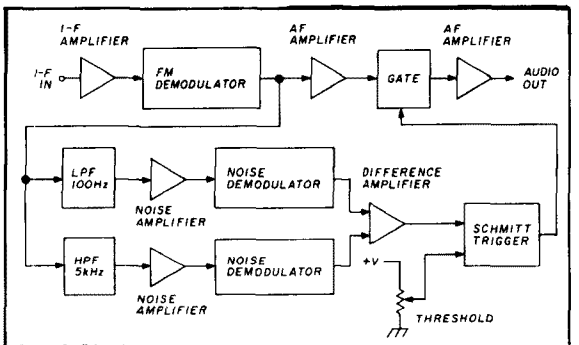


fig. 12. Block diagram of improved noise-operated squelch circuit for fm.

However, in some commercial communication systems, with multi-user net operation on one frequency, a coding scheme known as selective call (SelCall) has been devised so that users need receive only those messages directed toward them. In this type of signaling scheme, the caller sends a multitone or digital code at the beginning of the message to indicate the identity of the called party or parties. Only if the receiver code matches the transmitted code is the output gate enabled to transmit the message to the receiver user. This type of system may be used with

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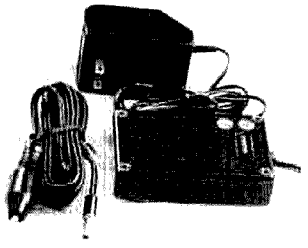
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both analog and digital modulations, and is independent of the modulation type (a-m, fm, pm) used for transmission. This method is more elaborate than normal squelch systems, but performs an important function in multi-user nets.

conclusion

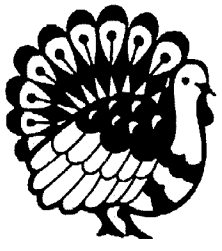
Several of the basic techniques and their circuits used to handle the introduction of noise in receiving systems have been described. While this discussion is by no means all-inclusive, it is nevertheless indicative of some of the basic approaches taken to solve this problem.

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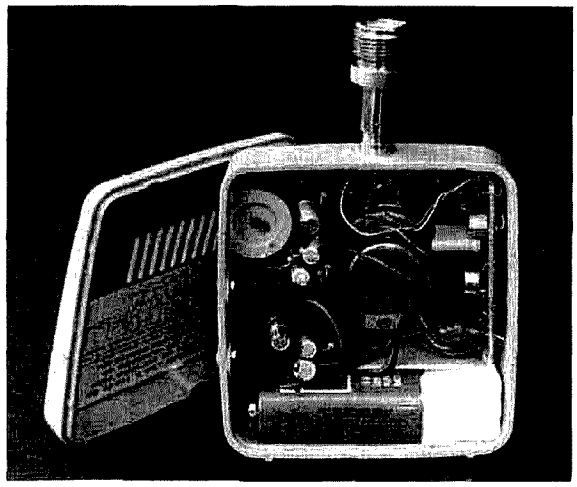
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a simple 80-meter receiver

Back in the good old days when AM reigned over the phone bands, many beginners built simple, one-tube receivers. As AM faded, however, so did the one-tube receiver; the more complex designs required for CW, SSB, and FM were beyond the skills of most beginners. This article describes an easy-to-build receiver for W1AW code practice and bulletins on 3.58 MHz, following a basic approach that minimizes construction difficulties, yet provides the satisfaction of building a useful accessory for even the most experienced Amateur.

Direct conversion receivers have been popular for some time now for CW and SSB. The complexities in the published designs for these units lie mainly in the tuning; the bandspreading and stability necessary to tune in SSB or separate crowded CW signals are difficult to achieve simply. Use of narrowband audio filters puts even tighter constraints on tuning.

Some time ago, I noticed that 3.5795-MHz TV colorburst crystals are in the 80-meter CW band, close to the 3.58-MHz frequency used by W1AW for code practice and bulletins; the difference in frequencies is a usable audio note. A crystal-controlled direct conversion receiver has sufficient stability for CW work, requires no tuning (because only one frequency is desired) or narrow filters (because 3.58 MHz is, relatively speaking, in the "wide open spaces"). The concept seemed useful, since a dedicated W1AW receiver would be a handy accessory as part of any code practice program. For areas where W1AW reception isn't reliable on 80 meters, other 80-meter crystals can be used for reception of other stations. Possibilities include W6OWP code practice on 3.59 MHz, which is heard on the West Coast, and perhaps local slow-speed CW net frequencies.

The receiver RF section fits on a small PC board which, in turn, fits nicely into the case of a small audio amplifier available from Radio Shack. This $12 off-the-shelf amplifier provides the case, more than adequate audio gain, a loudspeaker, earphone jack, volume control, and battery.

Although the receiver is battery-operated and fits easily in the palm of your hand, it does require a full-size antenna. For portable use, a pair of 25-foot long wires connected as a dipole should suffice. For stationary use, an 80-meter dipole (135 feet overall) would be better.

circuit description

The receiver shown in **fig. 1** is a direct conversion type with a preselector, mixer, and local oscillator. The preselector passes 80-meter signals and rejects all others. The mixer combines the received signal with

By Ed Gellender, WB2EAV, 28-37 212 Street, Bayside, New York 11360

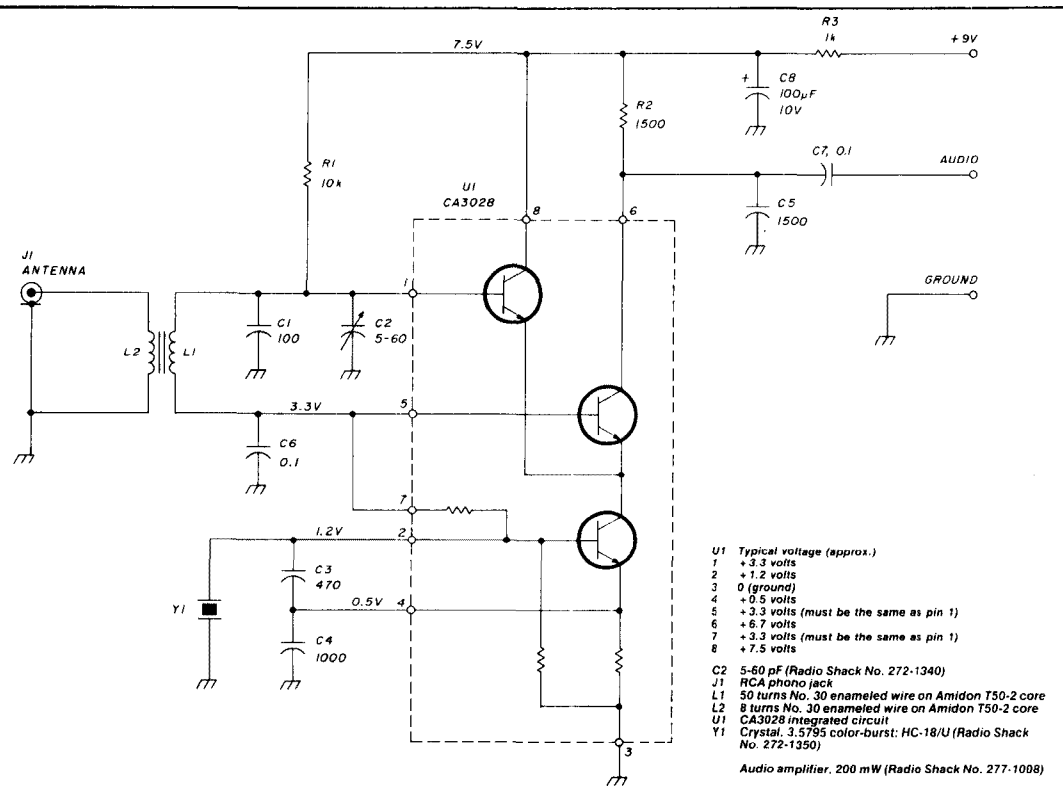


fig. 1. The heart of the simple 80-meter receiver is the CA3028 integrated circuit.

the oscillator output and provides an audible beat note, or difference frequency.

The ARRL Handbook shows several mixer circuits using the RCA CA3028 integrated circuit. The CA3028 consists of a differential amplifier and a common current source. All the published designs use the differential amplifier as a mixer and the current source transistor to buffer the local oscillator. Because we're using a crystal, the stability is good enough so that a buffer isn't needed, and we benefit from the simplification of using the current source transistor as the local oscillator.

The oscillator circuit is the classic Colpitts design, using phase-shift network C3, C4, and Y1 to feed signals from the emitter back to the base of the transistor. For the TV colorburst crystals specified to oscillate reliably, large values of C3 and C4 are required. As a result the crystal will oscillate just slighty above its series resonant frequency. If other crystals are used, their series resonant frequency should be about 1 to 1.5 kHz below the desired reception frequency.

Preselector L1, L2, C1, C2 resonates at 3.58 MHz with a loaded *Q* of approximately 15. C6 is an RF bypass capacitor that maintains U1 pin 5 at RF ground. The differential amplifier is driven by the RF input so that currents into pins 6 and 8 alternate at the input frequency. The total current shared by pins 6 and 8 varies with the local oscillator frequency. Nonlinearities in the base-emitter junctions result in a difference frequency appearing at pin 6. R2 and C5 form a low-pass filter to remove both the input and oscillator frequencies and pass the audio beat note. C7 is a DC blocking capacitor.

The audio is brought to the audio amplifier, which uses a single transistor preamplifier driving an LM386 amplifier IC. According to the amplifier data sheets, the amplifier sensitivity is 1 mv and maximum output is 200 mw. These specifications match the desired performance quite well. Purists might object to buying such a large part of the receiver ready-made, but practical considerations don't permit missing such an opportunity. As purchased, the amplifier's upper frequency response is about 10 kHz — too high for our purposes. But by changing one capacitor, we reduce the audio bandpass to a more usual 3 kHz.

constructing the RF board

The RF board is a 3/4-inch x 2-1/8-inch PC board, although a hand-wired board of those dimensions is suitable. The PC artwork is shown in **fig. 2** and the components layout in **fig. 3**. Be sure to install U1 as



fig. 2. Full-size pc pattern is shown from the DIP side.

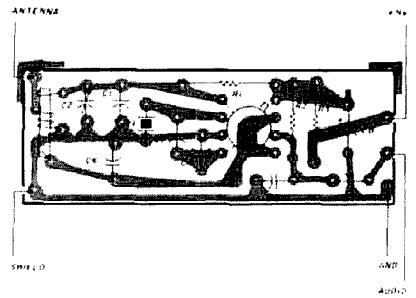


fig. 3. Recommended parts placement (component side).

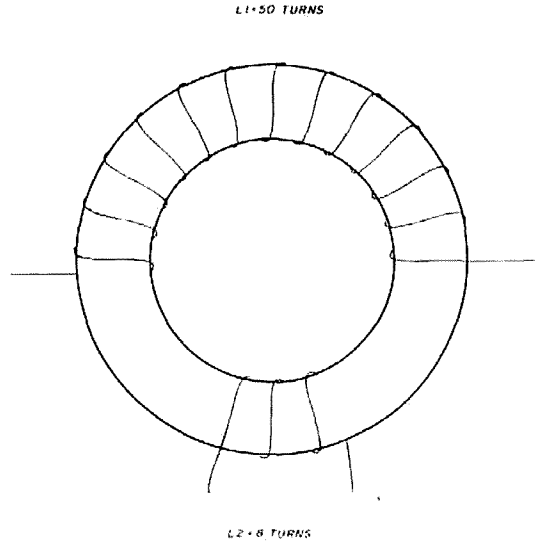


fig. 4. A simplified sketch illustrates the winding procedure for coils L1 and L2 on an Amidon toroidal core.

shown in the parts location diagram. The tab is over pin 8 (pins are numbered clockwise when viewed from the bottom). Bend the leads with a pair of needle-nose pliers to fit the hole pattern.

Coils L1 and L2 are wound on an Amidon Associates T50-2 ferrite toroid (see **fig. 4**). It must be wound as shown to fit the PC board pattern. To wind L1, take a 3-foot length of 30 gauge enameled magnet wire and tightly wind 50 turns. This should occupy about 2/3 of the core and use about 30 inches of wire. Don't forget to leave pigtail leads on each end. Similarly wind L2 in the remaining empty part of the core. Before soldering, scrape the insulation off the wire pigtail leads with a razor blade. A few drops of glue (such as Duco Cement™) can later be used to hold it to the board.

The crystal specified, an HC-18/U type, is soldered to the board. The miniature crystal lends itself well to this construction, but if one wishes to use other crystal types, there's no reason short wires couldn't be used to connect a larger crystal or a crystal socket. If other crystals are used, however, the values of C3 and C4 might have to be changed for the oscillator to start reliably.

If a single small package isn't necessary, or if you want to breadboard the RF board differently, a separate RF breadboard can be used (see **fig. 5**), with the audio output brought to a miniature phone plug that fits into the audio amplifier input jack. If separate 9-volt batteries are used, the audio amplifier used should not be modified (except for adding a recommended 0.01 μF capacitor described below).

modifying the amplifier

The Radio Shack Audio Amplifier must be modified slightly before the RF section is added. First, lift the audio amplifier PC board out of the cabinet. To do this, unscrew the two knurled nuts from the outside of the two miniature phone jacks and unscrew the small phillips-head screw at the top of the unit. Then, while applying finger pressure to "belly-out" the plastic case away from the phone jacks, lift the board from the loudspeaker side.

To install the antenna jack, carefully drill a ¼-inch hole through the top center of the cabinet and install J1, an RCA-type phono jack. Be sure to use the ground lug.

The input jack is of the shorting type and must be modified. Unsoldering the jack and removing it is one possibility. Another possibility is to cut the land between the center terminal of the input jack and the volume control. A third possibility, which allows the amplifier to be used for other purposes, is simply to leave an open-circuited miniature phone plug in the input jack. Solder an insulated wire 3 or 4 inches long to the land between the volume control and input jack, on the volume control side of any out lands. This is the audio connection lead and is soldered to the pad on the RF board connected to C7.



fig. 5. Few leads are needed to interconnect the rf board to the rest of the circuit.

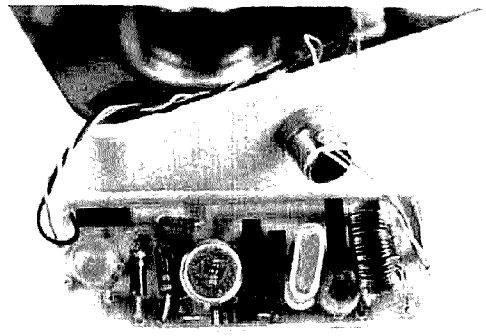


fig. 6. Sufficient space is available in the Radio Shack audio amplifier for mounting the rf board.

One optional but recommended modification is to remove the 0.0033 μF capacitor located in front of the small plastic transistor and replace it with a 0.01 μF ceramic capacitor. This reduces the audio frequency response to about 3 kHz, eliminating much noise.

Now add the remaining wires to the RF board. Connect a ground wire from the large ground plane in the center of the audio board to the ground pad on the RF board near C7. Connect another ground wire from the ground lug on the phono jack at the top of the cabinet to the ground pad at L2. Connect a wire for the switched +9 VDC from the bare jumper between the two phono jacks to the 9-volt pad on the RF board, near C8. Finally, connect a wire from the center conductor of the phono jack on the top of the cabinet to the antenna pad at L2 (see fig. 6).

Reinstall the amplifier board in the case (it goes in more easily than it came out). Place the RF board, vertically, in the open space on the other side of the loudspeaker. L1 should be near J1 and C8 near the battery. A grommet or suitable non-conductive spacer placed between the top of U1 and the side of the loudspeaker will hold the RF board in place.

operation

Connect a 9-volt battery to the battery clip in the amplifier. Turn on the amplifier with the volume control. Check the pins of U1 for the DC voltages shown on the voltage chart. The amplifier should output a low hissing noise from the speaker with the volume turned up. If another receiver is available, listen for the oscillator signal at 3.58 MHz.

Connect the receiver to a suitable antenna. The background noise should increase noticeably as the atmospheric noise in the 80-meter band overrides the internal noise. Signals may be heard. Variable capacitor C2 is set for strongest reception of the desired signal. Note that for some antennas, the best adjustment of C2 may cause the mixer stage to oscillate when the antenna is removed, as indicated by a raspy signal. If oscillations occur while the antenna is connected, slightly detuning C2 is all that's necessary.

Private listening is possible, using an earphone with a 1/8-inch phone plug; these are often supplied with broadcast radios, but are also available separately.

performance

The receiver is capable of detecting signals far below the atmospheric noise level in the 80-meter band, so sensitivity is not a problem. With strong signals, room-filling volume is available from the loudspeaker. There are no controls other than the volume control. Because of the lack of tuning and filtering, interference from nearby signals can be a problem. However, in the "wide open spaces" at 3.58 MHz, this won't be encountered often.

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receiver tuning mechanism selection

What tuning system to use? Read on.

One of the critical steps in designing a communications receiver involves deciding what tuning system will be used. Part of this decision includes determining the tuning rate (kHz covered per turn of tuning knob), band coverage (kHz covered over an entire band), and tuning direction.*

Several related details must also be considered. For example, in selecting the tuning rate, it's also necessary to choose the appropriate knob; a crankshaft knob generally offers smoother tuning than a round knob of the same diameter. Similarly, the band coverage selected will be related to the mechanism reduction ratio and to the tuning rate.

Because most modern receivers have digital readouts, the probability of ending up with opposite-direction dial scales (as a consequence of conversion schemes) is slim, but is something to remember in case analog dial readouts are used.

tuning mechanisms

It's necessary to decide, early in the design stage, whether tuning will be via permeability or capacitance variation.

Permeability tuning offers the advantage of no sliding electrical contacts. It's probably the easiest system to build if you don't have an elaborate workshop. Inductance variation is achieved by moving a ferrite core in and out of a coil of large length-to-diameter ratio, generally 5:1 or more. The best linearity over the frequency range occurs when the core is about 30 to 40 percent inside the coil. Movement consists of the rotation of a threaded shaft, to which one or more springs are attached to eliminate thread backlash and lateral movement. By simply selecting a particular thread pitch and designing the coil winding accordingly, frequency coverage and tuning rate can be designed to meet specific requirements. Very often a combination of series and parallel coils is used to achieve the desired coverage.

Using a 6-32 threaded shaft and a 1/4-inch diameter, 1-inch long coil, one can easily build a VFO into a 1 x 1 x 2-inch box. **Figure 1** shows the basic construction; the VFO printed circuit can be fitted below the coil, fastened to the 1/8-inch aluminum bracket housing the VFO itself.

Construction details for a 5- to 6-MHz permeability tuned VFO is shown in reference 1.

Multicoil setups used in receiver front ends can also be built using the same concept, even though mechanical difficulties tend to increase as the number of coils increases. Regardless of the number of coils, end stops must be used at both coil ends to avoid damaging the mechanism or having the ferrite core fall off the threaded shaft.

An alternative method of permeability tuning (**fig. 2**) is spring-loading the ferrite core against a cam moved by the tuning shaft. In pre-digital days, this technique allowed one to linearize the VFO by filing the cam profile rather than by adjusting the coil winding.

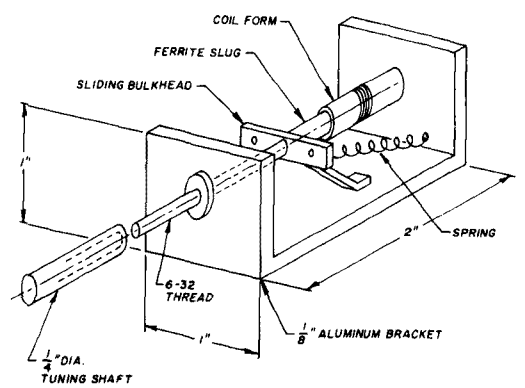


fig. 1. Permeability-tuned VFO (end stops not shown).

*An unwritten convention prevails: clockwise rotation increases frequency. In some British and Japanese sets, however, counterclockwise rotation increases frequency.

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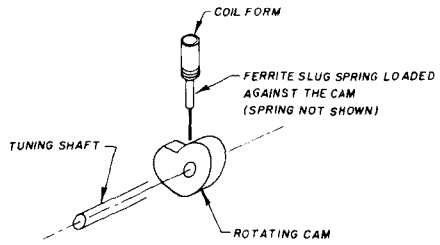


fig. 2. Basic setup of cam permeability-tuned VFO.

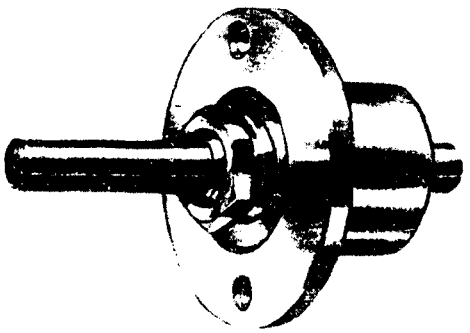


fig. 3. 10:1 ratio heavy-duty ball vernier by Jackson Brothers (UK). This unit has output torque exceeding 30 oz/inch. The front nut fastens a rotary dial.

With digital readout receivers, this method offers no substantial advantage over the standard mechanism shown in **fig. 1**, except for the simplicity inherent in determining the two end stops. Otherwise it's more mechanically complex.

Capacitance variation is possibly the most widely used system for receiver tuning.

The drum and chord is standard in older receivers, and many portable units. Readily-available, low-cost drums and variable capacitors with bearings at both ends make this one of the easiest tuning mechanisms to homebrew. Results are excellent, particularly if stainless steel chord, rather than plastic or synthetics are used. One convenient feature of this system is its ability to slip at both tuning ends, *thus providing a built-in safety* device to avoid damage to the variable capacitor. One disadvantage of the method, however, is that the reduction ratio is somewhat limited, seldom exceeding 30:1 or 40:1 over 360 degrees.

Many types of *vernier reducers* are available. These are primarily based on a row of steel balls rotating on an inner and outer machined groove. The difference in diameter between the inner and outer ball races provides the reduction effect, which seldom exceeds 10:1. Jackson Brothers, a British manufacturer with distributors in the United States, offers an extensive line of verniers; a good unit to use is their Model 5857, depicted in **fig. 3.**

Some manufacturers build the vernier right into the variable capacitor shaft, making the whole tuning mechanism quite compact, though the reduction ratio is low. In both cases the reduction ratio is generally too small for serious communications work. For instance, if one does not want to cover more than 50 kHz per turn (remember, the full capacitance swing of a variable takes place in about 160 degrees) a vernier with a 10:1 ratio will offer only 160 degrees/360 degrees times 10 turns, or approximately 4.4 turns for the total swing of a standard capacitor, thus limiting the band coverage to 4.4 turns times 50 kHz per turn, or 220 kHz.

Another limitation is aging. With aging, the vernier tends to slip or to develop backlash, especially if the variable capacitor in use requires high torque to rotate.

Gear reduction, widely used in commercial and professional equipment, can be of the parallel or worm

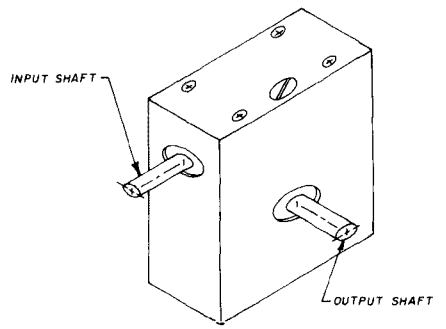


fig. 4A. Typical single worm gear reducer. The input shaft is mounted in precision ball races; the output shaft, in self-lubricating bearings. The final assembly is packed in grease having a temperature range of −34 to +150 C.

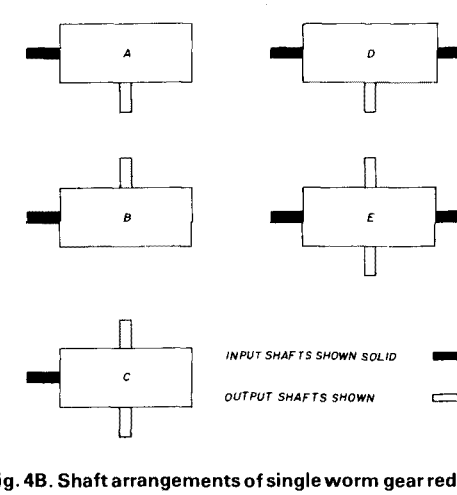


fig. 4B. Shaft arrangements of single worm gear reducers.

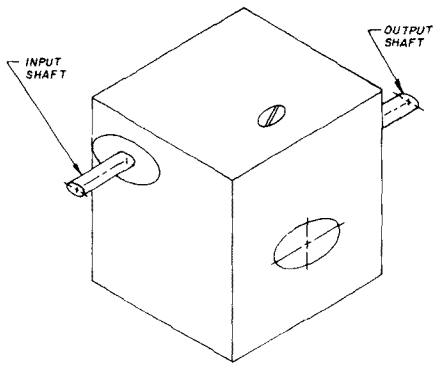


fig. 5. Sealed zero-backlash double worm gear reducer manufactured by Muffett Ltd. (UK).

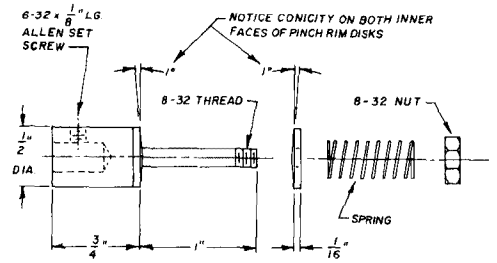


fig. 6. Pinch rim drive can be home-built from aluminum or brass round stock. The spring should be made of spring wire measuring approximately 0.025 inches in diameter and wound in five to eight turns on a 3/16-inch diameter drill.

gear variety. In both cases, the gears must be spring-loaded to eliminate backlash. Both systems can offer any reduction ratio but, for the same reduction ratio, parallel gear reducers tend to be bulkier. From an Amateur standpoint, the availability of parallel gear reducers is more limited, and new reducers tend to be expensive — i.e., in the range of $50 to $80.

A specialized parallel gear reducer is an epicyclic drive, in which the outer (larger) gear is an "internal" gear. (This means that the teeth protrude inward, rather than outward, as on standard gears.) Its most popular form is the unit used in the National HRO series. The epicyclic drive can be home-built or purchased from a manufacturer of precision gears for between $75 and $100.

One limitation of this system is its low reduction ratio, which seldom exceeds 15:1 for 360 degrees rotation. A wide variety of worm gear reducers with ratios in the range 20:1 to 100:1 is available on the surplus market; an entire set can be bought for just $10 or $15. However, because these units are quite old, the tuning mechanism may show some wear and may have developed backlash. Furthermore, by today's standards, these mechanisms are generally heavy and bulky, and will require mechanical adaptation if they're to be used in a modern project.

Used extensively in industrial instrumentation and aerospace applications, new worm gear reducers are available from several suppliers. Versions with ball bearings on both input and output shafts, housed in sealed, lubricated-for-life assemblies, offer a truly superb, "velvet" feeling (**figs. 4A, 4B**). Made by Muffett, Ltd. (UK), they're distributed in the United States by Sterling Instruments of New Hyde Park, New York. The size most compatible with ham gear is No. 2, which is offered in reduction ratios up to 100:1, at about $75.

Another interesting item is a double worm gear reducer, which consists of two reducers cascaded inside the same assembly, thus offering very high reduction ratios (up to 10,000:1) that may be used for specialized applications (**fig. 5**). On all worm gear reducers, because of their high reduction ratio, means must be provided to have end stops at both ends of the variable capacitor excursion to avoid damage.

Pinch-rim drives. In this method, a disc is mounted on the variable shaft and the tuning shaft pinches laterally the periphery of the disc with two spring-loaded discs of much smaller diameter. The system slips automatically at the variable end stops, providing a safety feature at both ends. It can be home-built with limited mechanical means, "turning" it on a hand drill fastened to a bench vise and making the main disc out of thin aluminum sheet or plastic. (See **fig. 6** and reference 2.) As in the case of the ball vernier, however, a limitation is the reduction ratio, which seldom exceeds 10:1 or 15:1 over a 360-degree rotation.

This system is sometimes cascaded with a gear reduction system that provides a clutching (slipping) action at both ends, resulting in "velvet" tuning.

conclusion

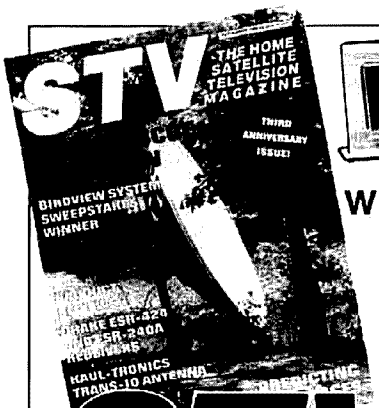
Early in the design of any communication gear, the homebuilder must decide on whether the tuning system will be of the variable L or C variety. While variable L is probably easier to build, the system is not as popular as it could be, mainly because even though the mechanical work required is limited, one can find a commercial vernier and variable capacitor rather easily these days. One word of caution: it's practically impossible to build a zero-backlash, high-frequency VFO with variable C unless the capacitor shaft has bearings at both ends. Standard trimmers or transmitting variable capacitors are bushing-mounted and generally show backlash in one of two ways. Either the rotor bushings are quite tight (to avoid the shaft fitting loosely in the bushing), with backlash developing in the tuning system as the torque required to rotate the shaft exceeds the rating of the reducer, or the rotor bushings aren't tight. In the latter

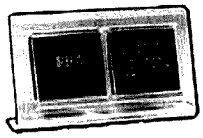
case, backlash develops within the variable capacitor itself because of the rotor's wobbling in and out of the stator. I take it as a rule that for decent communications work to be performed, the variable capacitor *must* have bearings at both ends.

references

1. J. Perolo, PY2PE1C, "A Solid State Permeability Tuned VFO with Digital Frequency Read out," *CQ*, October 1970, page 18.
2. J. Perolo, PY2PE1C, "Portable Short wave Receiver," *ham radio*, August, 1984, page 67.

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Achieve better noise floor and "crunch-proof" performance

upgrading the Ten-Tec Argosy

This article proves two things: first, that outstanding results can be obtained by choosing inexpensive, well-designed equipment and second, that there's still room for hams to make worthwhile improvements to their transceivers. Owners of the original Ten-Tec Argosy who'd like to hear weak signals in the presence of powerful adjacent signals without "popping" and intermodulation distortion may find it especially useful.

With several thousand in operation around the world, the Argosy is particularly attractive for the CW QRP operator. It can also serve as a low-cost base station transceiver which can drive a high-gain linear amplifier to 1000 watts output on most bands.

Although my Argosy was a good performer in many respects, I decided that it could be improved. For example, with DX propagation only marginal (given the current sunspot cycle), its noise floor was so high that DX signals of S3 (approximately 0.8 μV rms in a 50-ohm system) and below were obscured. In addition, the receiver desensitized 100 kHz away from a signal generated by a station located no more than a mile from my QTH.

The modifications described here will improve the noise floor, compression point, intercept point, and AGC characteristics of the Argosy receiver, as well as the transmitted characteristics of its SSB signal. They will also add digital versatility to the unit by replacing the analog dial with a precise four-digit display capable of reading the transceiver's frequency with an accuracy of ± 10 Hz.

Critical listening and laboratory tests comparing a modified Argosy with an unmodified one have shown that these modifications result in significant improvement (**fig. 1**). The test results (**table 1**) show clear improvement in the areas of minimum discernible signal (MDS); two-tone, spurious-free dynamic range (SFDR); and blocking dynamic range at only 5 kHz from the interfering signal, with the latter showing an impressive 30-dB improvement over the unmodified Argosy. After the modification, on-the-air tests allowed copy of weak CW DX stations of S3 and below as close as 2.5 kHz from WB0NHD, less than a mile from my QTH and running a full kilowatt with his antenna pointed in my direction!

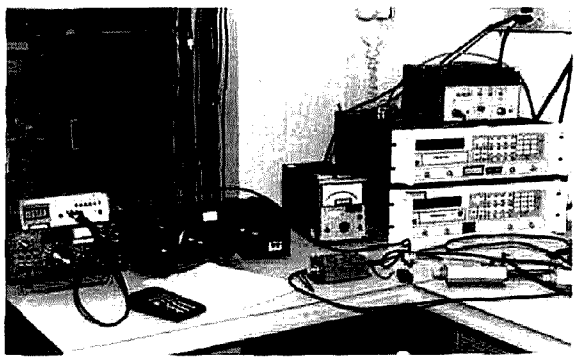


fig. 1. Laboratory setup used for testing the dynamic range of the two Argosys. The tests were performed with two tones at 14.020 and 14.040 MHz. Third-order product was recorded at 14.060 with the help of a true RMS voltmeter.

By Cornell Drentea, WB3JZO, 7140 Colorado Avenue North, Minneapolis, Minnesota 55429.

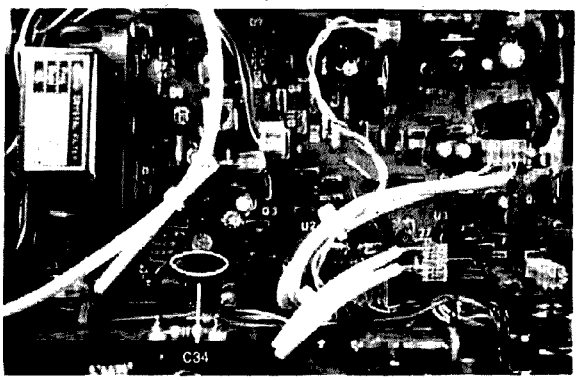


fig. 2. Replacing C34 on the IF/AF board improves the minimum discernible signal (MDS) of the receiver. See text.

While some of these changes were communicated to the manufacturer and reportedly incorporated in the production of the recently-released *Digital Argosy*, owners of this unit may wish to perform those modifications not implemented by the manufacturer.

The changes described in this article are within the reach of any technically inclined Amateur and can be implemented totally or partially as desired. However, I recommend that you have a good understanding of how the Argosy works before proceeding with the modifications. I've tried to describe the modifications in sufficient detail here; because of limited time, neither Ten-Tec nor I will be able to provide additional information. For a better understanding of this article, it would be wise to read the owner's manual as carefully as possible.

improving the noise floor

As previously stated, one of the most annoying problems associated with my Argosy's receiver performance was its high noise floor that obscured weak stations. While this was not apparent on strong signals, the hiss greatly reduced the overall intelligibility of many CW and SSB signals.

Several calls and a face-to-face discussion at the Dayton Hamvention with Ten-Tec engineers offered some possible solutions. After several unsuccessful trials, I proceeded on my own to find out what was wrong with the receiver. A few hours of analysis and careful isolation of stages revealed, much to my surprise, that the problem was caused by a design glitch in a totally different area of the receiver. The audio low-pass filter (R47, C34 following the product detector, Q3, which is located on the i-f/af board 80785) had a corner frequency of 3.28 kHz, which would have been a good choice if attenuation were steep beyond this point. However, this was only a simple first-order low-pass filter with an attenuation slope of 6 dB per octave, which was incapable of preventing BFO noise sidebands of up to 10 kHz from spilling into the high gain audio amplifier of the receiver after being mixed down to audio frequencies in the product detector. This, in turn, superimposed an annoying hiss over all the received signals and obscured signals below S3. Once the problem was found, the fix was simple. Changing C34 from 0.01 to 0.1 started the corner frequency of the low-pass filter at 338 Hz, allowing enough roll-off beyond 3 kHz, cancelling the annoying noise and making a dramatic improvement in the signal-to-noise performance of the receiver. Although one might suspect that the 6-dB roll-off of the new filter with a corner frequency of only 338 Hz might affect the overall audio response of the receiver, this was not the case. The audio frequency response measurements performed after the modification was made showed a flat response over the entire audio range of interest (600 Hz to 2800 Hz).

Table 1: Comparative dynamic range tests between an unmodified and a modified Argosy were performed under laboratory conditions. Tests were performed in the 14-MHz CW band with all crystal and audio filters "in", and the noise blanker out of the circuit.

Spec	Unmodified	Modified
MDS	– 126 dBm	– 138 dBm
Blocking dynamic range, 5 kHz from the interfering carrier	103 dB	133 dB
Spurious-free dynamic range (SFDR) two-tone, third-order 20 kHz spacing	80 dB	98.5 dB

This was due to compensation in the audio gain of the receiver inherent in the original design. With the problem fixed, the rig worked so well that I decided to share the information with Ten-Tec, whose engineers quickly calculated the new corner frequency, tested the change, and concluded that it made a "radical" difference in the noise floor performance of the receiver.

Figure 2 shows the location of C34 in the transceiver. While it's not necessary to completely remove this capacitor (one can parallel the new capacitor on top of the board), doing so involves simply removing the board — just unplug all connectors, remove all metal screws, and replace C34 with a 0.1 μF ceramic capacitor rated at a voltage greater than 25 volts. Slide the board back into place. This modification should take no more than 10 minutes. (Be careful to plug connectors back exactly where they belong.)

Even better performance can be obtained by totally replacing the RC network with a second-order filter with a cutoff frequency of 3 kHz. This involves the use of a high-Q, 470-mH inductor as shown in **fig. 3**. In this modification, the source resistor (4.7 k) has been

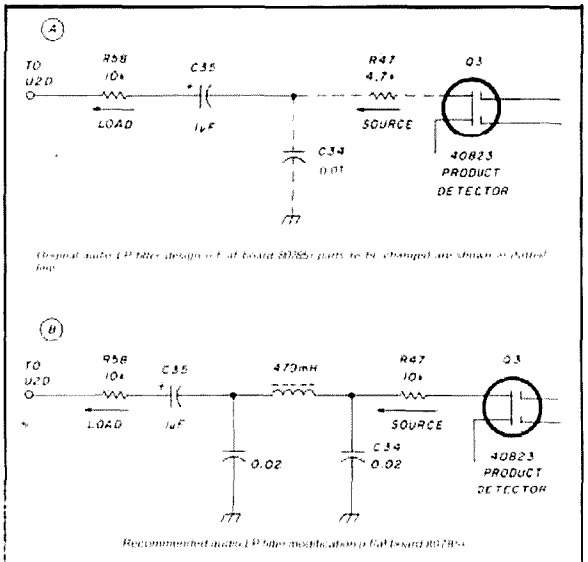


fig. 3. Further improvement of the minimum discernible signal (MDS) can be obtained by replacing the first order RC low-pass filter (R47, C34) with a second-order high Q inductor type filter with a cutoff frequency of 3 kHz. The 470 mH ferrite inductor is available from the author.

replaced with a 10-k resistor, which was necessary for symmetrical C values in the legs of the filter, as well as to make use of the entire range of the volume control.

reducing ac hum

Although the receiver was now very quiet, it was immediately apparent that the next problem was a 120-Hz hum previously obscured by the noise floor. This hum was caused by the ac switch and power line located in the immediate vicinity of the high-gain audio stages, unshielded audio cables, and front panel controls, all acting as audiq pickups for the induced ac field.

I decided that the perfect solution — eliminating the power switch at this location — would not only solve this problem but also allow another modification: the insertion of an rf/af gain, dual potentiometer, which will be described later in this article. Good results can be obtained by shielding both the power line and the sensitive cables as shown in **fig. 4**. Copper foil is recommended; however, aluminum foil should also work. Make sure that the shield is grounded at several points to avoid ground loops.

improving dynamic range

Two additional modifications were effective in increasing the Argosy's dynamic range. While they are specifically intended to improve the intercept point of the receiver, they also improve MDS performance.

The first modification consists of replacing the present mixer diode set (D12, D13, D14, and D15* on the rf/mixer board 80784) with a new set of matched high-voltage Schottky barrier diodes. The part chosen for this application was the Hewlett Packard 5082-2800. The 2800 uses a "guard ring" design, making it ideally suited for handling high-level signals.

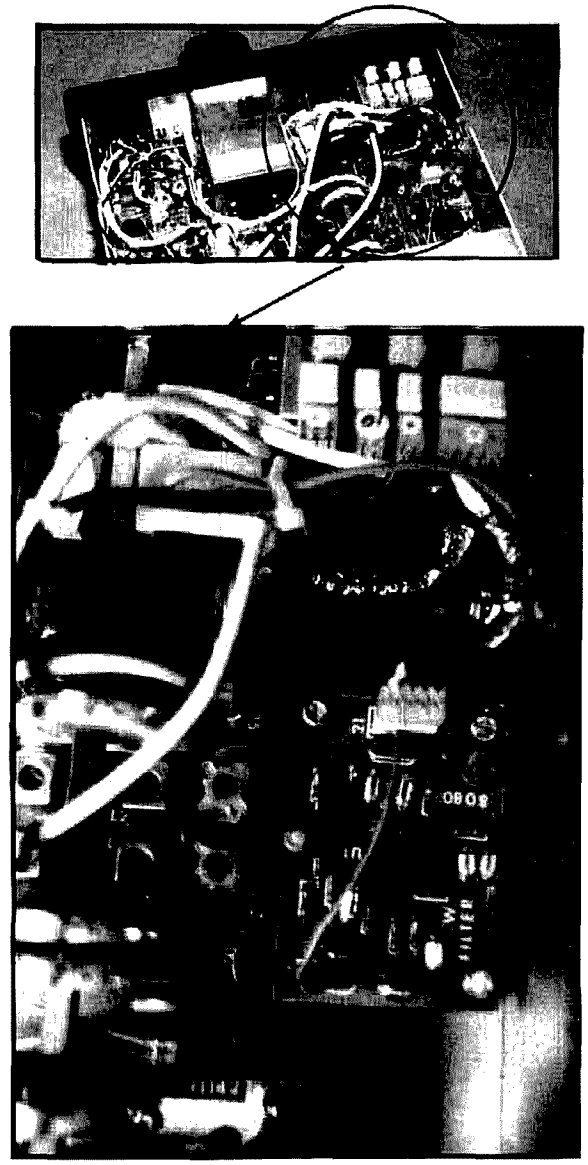


fig. 4. Shielding the power line and sensitive cables in the high gain af section of the Argosy will provide further signal to noise improvements in the receiver. (A) Bottom view of chassis. (B) Expanded view of cable assemblies.

The second modification consists of changing the rf switch diodes (D10 and D11) on the same board to high-

*_ham radio_ style customarily identifies diodes as "CR." To avoid confusion, however, the "D" nomenclature is used here when referring to Argosy components.

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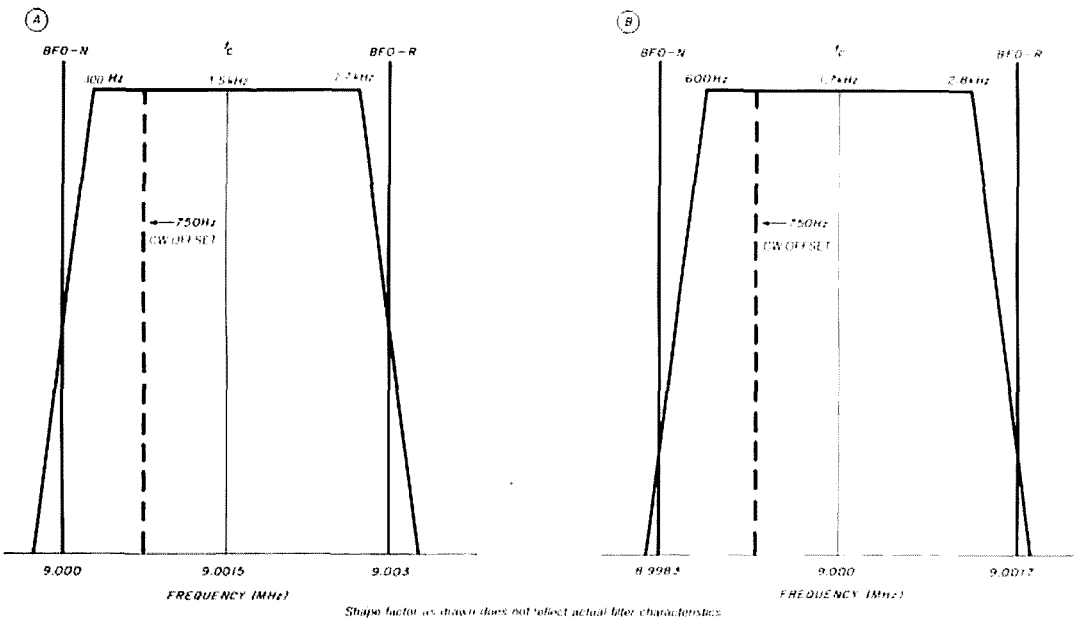
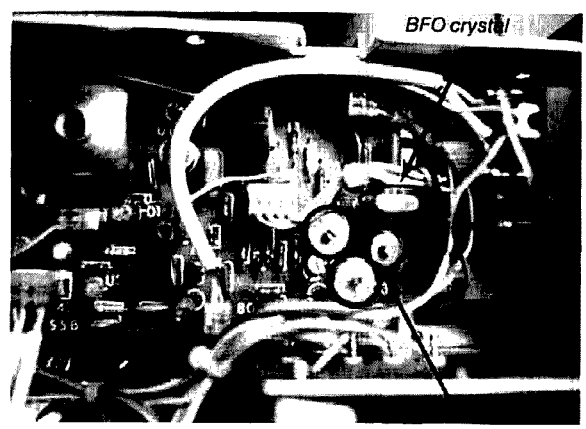


fig. 5. (A) BFO adjustments for the original 9.0015-MHz Argosy filters; (B) BFO adjustments with standard 9.000-MHz filters.

current PIN diodes designed especially for rf switches and attenuators such as the HP 5082-3081. About 75 mA of bias current should be provided to these two diodes for a positive rf turn-on. This is accomplished by replacing the 470-ohm resistor at R4 with a 220-ohm half-watt resistor. Although the 3081 can use more current for better rf switching, this is not recommended in order to protect the drive transistors located on the Control Board 80781. This modification is not recommended for applications requiring thrifty battery use — for example, in portable QRP operation. However, it does further improve the dynamic range of the receiver.

To implement these changes, remove all connectors and metal screws carefully and slide the rf/mixer board out. Remove the old diodes with the help of "Solder Wik" or a similar product and replace them with the new parts, following the diode polarity already marked on the board (the band is the cathode). Reinstall the board, making sure that all metal screws and connectors are back in their proper places. Readjust mixer balance trimmer C2 for minimum received product at 21.320 MHz, as outlined in the manufacturer's manual (pages 3-7). The receiver will now be very quiet and signals will stand out against the background without being disturbed by powerful adjacent stations. The AGC will also be favorably affected since less intermodulation distortion will be present in the i-f passband. Remarkable results have been realized by replacing the original four-pole i-f filter with a better filter, but some problems will occur if you try to use standard 9.000-MHz filters.



C_1, C_2, C_3 settings with new BFO frequencies

fig. 6. New trimmer settings required for the SSB generator Board 89789 when using the recommended 9.000 MHz filters.

using standard 9.000-MHz i-f filters

The Argosy came equipped with a four-pole, 2.4-kHz wide filter which drew some unfavorable audio reports in SSB. In addition, a "wide" transmitted signal was reported on some crowded bands.

A similar problem noted in the receive mode contributed to adjacent signals getting into the i-f passband and consequently activating the AGC. While it's possible to buy optional filters from the manufacturer, this was be-

yond my financial means at the time. I tried some new 9.000-MHz filters with the proper impedance but found they produced unfavorable results for no apparent reason. A careful analysis of the design showed that the manufacturer's choice of i-f center frequency was 9.0015 MHz instead of the more popular 9.000-MHz approach (this isn't readily apparent from either the manual or the part numbers marked on the filters as used in the transceiver). Because 9.0015-MHz filters were not available to me, some ingenuity was required in order to use the more standard 9.000-MHz filters.

In order to use these filters with the Argosy it was necessary to change the frequency of the BFO crystal and adjust three trimmer capacitors on the SSB Generator Board 80780 for slightly different frequencies.

Let's analyze the way the BFO works from **fig. 5A**. In order to use the 9.0015 center frequency filter, the Argosy designers chose BFO frequencies of 9.000 MHz for "SB-Normal" and 9.003 MHz for "SB-Reverse." These parameters are set by frequency-pulling a crystal oscillator via the triple trimmer setup on the SSB generator board according to a procedure outlined in the owner's manual.

To use 9.000-MHz filters, I replaced the BFO crystal with an 8.9985-MHz crystal (a KVG XF-901) and proceeded to recalibrate the three trimmer capacitors according to the Carrier Oscillator Alignment Procedure explained in the manual and by using the new parameters shown in **fig. 5B**. The new numbers allowed for an audio passband of 600 Hz to 2800 Hz (center frequency, *fc*, is 1700 Hz) when using the filters recommended in this article (see table 2). The 1700-Hz center frequency was selected over the more conventional 1500-Hz unit after a series of intelligibility tests. Many excellent reports were received after the modification.

This audio range lends itself easily to the 750-Hz offset required for the CW mode. The recommended 9.000-MHz filters exhibit superior shape factors over the original four-pole filter (See **table 2**). **Figure 6** shows the approximate position of C1, C2, and C3 trimmer capacitors necessary for this modification as implemented in my transceiver.

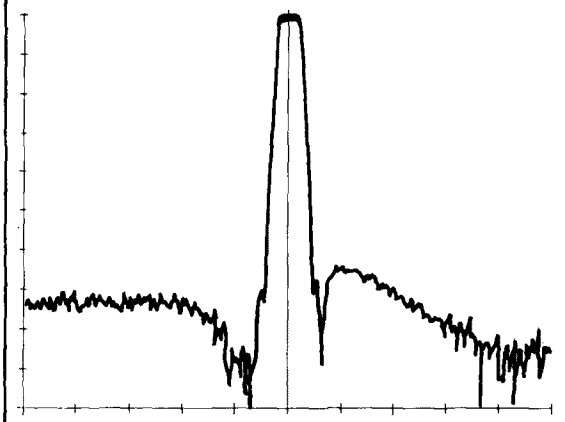
A set of two 9.000-MHz filters can be cascaded (one replacing the four-pole filter on the rf/mixer board 80784 and the other at the optional location of the i-f/af board 80785) with exceptional results in receive or transmit. The combination is the equivalent of a 2.2-kHz wide, 12-pole filter with a shape factor of better than 1.6:1 in the receive mode (only one filter is used in transmit). When combined with the previous changes, this modification produced superlative performance which included improved AGC response in all modes and improved small signal performance in the presence of strong signals.

To make the change, construct the filter connectors by cutting two 2.4 x 1.0-inch pieces of unclad PC board material. Drill four holes to fit four 0.040-inch diameter connector pins snugly. (These pins can be made of discarded TO-5 transistor leads.) The connector pattern can be copied directly from the transceiver PC board layout with a piece of paper and a pencil, using the pencil to punch the existing connector holes and transferring the information to the new PC boards. Drill additional holes for installing the filter assembly on the PC board as required. Then secure the four connecting pins in place at the proper height with blobs of solder on each side of the board, following the dimensions from the original filter assembly. Connect the inputs, outputs, and the grounds of the filters in the same manner as in the original filter. It's important to realize that the cascaded filter on the i-f/af board 80785 will require two dc blocking capacitors if you're using the suggested filters (0.01 μF will do) at the input and output ports because of the diode switching involved there.

After installing the filters and the new BFO crystal, perform the manufacturer's carrier oscillator alignment on the SSB generator board 80780 by using 8.9983 MHz for "SB-Normal" and 9.0017 for the "SB-Reverse" positions. Use **fig. 6** as a guide to begin the setup and fine

Table 2: Specifications of the 9.000-MHz filter used in the modification.

Center frequency	9.000 MHz
Bandwidth at 6 dB points	2.20 kHz
Shape factor, 60-6 dB	1.8
Source impedance	500 ohms
Load impedance	500 ohms
Number of poles	6 crystal poles; 3 LC poles
Insertion loss	Less than 6 dB
USB BFO setting	8998.3 kHz (or 8998.5)
LSB BFO setting	9001.7 kHz (or 9001.5)



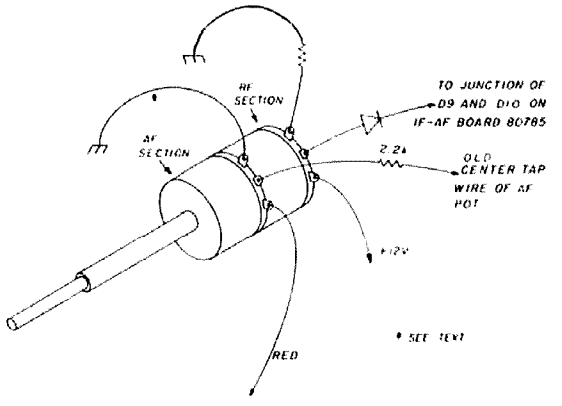


fig. 7. Wiring a dual concentric 10 k potentiometer for rf and af gain controls in the Argosy. The introduction of the rf gain allows control over the AGC when copying weak stations in the presence of strong ones. This modification completely eliminates the popping phenomenon.

tuning of the three capacitors until no error results from the interactive effect (this modification requires a full understanding of how the Argosy BFO works). Adjust the CW offset in step 10 of the procedure to 750 Hz above the "SB-Normal" as required. Although the procedure outlined in the owner's manual is a bit confusing at first because of the interaction between the three capacitors, using a digital counter and a scope should make things easier. No matter how well the settings are performed, the BFO frequency will change value slightly over time because of environmental impact on the trimmer capacitors. Variations of plus or minus several tens of Hz are not unusual for this BFO. These variations can introduce some coloration when you're switching sidebands in either the transmit or receive modes.

To check whether the transceiver needs realignment, follow this procedure. With the receiver on and the antenna disconnected, listen for the "hiss" pitch characteristic in SB-N. If things are "equal" (which they should be), the same pitch characteristic should be heard when switching from SB-N to SB-R. A final test for a different kind of coloration (called *SSB brilliance*) should be performed in the transmit mode with the rf output fed to a dummy load and a microphone connected to the transceiver. To verify that your SSB signal sounds "alive" on the air, use another receiver equipped with a pair of headphones to listen to your own signal. To avoid overload of the receiver, perform the test in the 5-watt QRP transmitter position. If the signal sounds mushy, realign transformer T-1 (the big can) located on the same SSB generator board while pronouncing words containing the letters "S" and "Z" into the microphone.

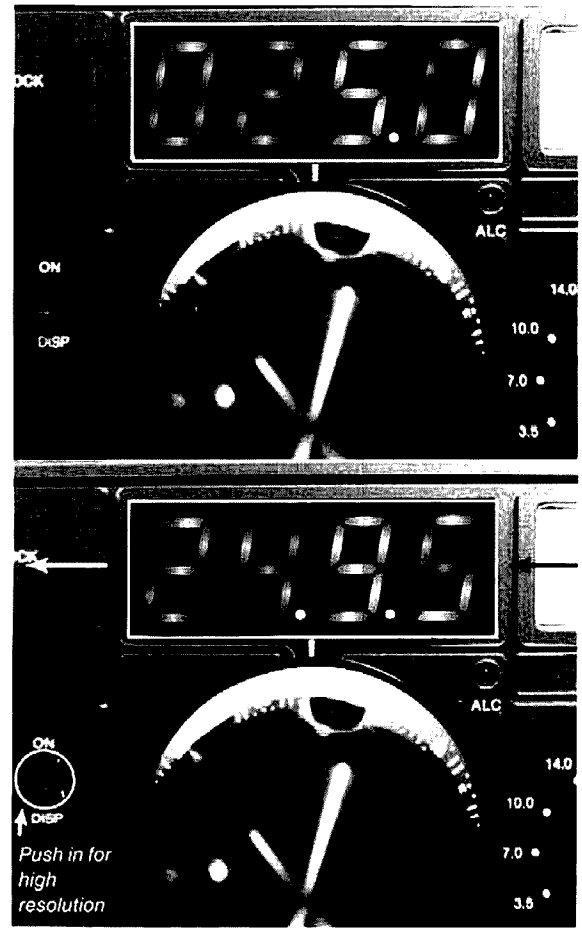


fig. 8. The Argosy frequency can be read to ± 10 Hz (short term stability of the rig is fully compatible with this feature) by pushing in the "CAL" or "DISP" button on the front panel red windows. An additional decimal point will light up in the display (see lower view) indicating that the number has shifted to the left.

Using the plastic tool provided with the transceiver, slightly vary the position of the T-1 core until these letters are heard clearly in the monitoring receiver (the core should be at the bottom of its travel). Repeat this test in the opposite sideband until everything sounds as lifelike as possible. Because the transformer has been adjusted at the factory, this adjustment should be minor. If it hasn't been properly set, this can make a major difference in intelligibility of your SSB signals at the other end. (It has no effect on the receiver itself.) Last, but not least, make sure that carrier balance adjustment, R1, is set for minimum carrier leakage as pointed out in the procedure described in the manual.

improving audio AGC characteristics

One of my biggest complaints was the popping

characteristic of the Argosy's audio AGC. While most of the modifications presented so far improve to a great degree the AGC's reaction to unwanted out-of-band signals and in-band background noise, a simple additional modification will virtually eliminate popping. This change consists of replacing C12 on the i-f/af board 80785 with a 22-μF, 35-VDC tantalum capacitor, which exhibits a favorable charging curve that cancels out the annoying "pop," at the same time providing a more suitable response time for the AGC. With this capacitor in place, the attack and release times were measured at 20 milliseconds and 2 seconds, respectively. This constitutes an almost perfect compromise between CW and SSB. (This modification can be incorporated when the other changes are being made to the same board.)

When installing the tantalum capacitor, be sure to observe polarity. After this change, popping was almost completely eliminated with only a hint of it left on signals exceeding 40 dB over S9 in level, and with all filters (including narrow af) inserted. In order to completely eliminate the popping, the addition of the rf gain control is required.

rf gain control

Although the original Argosy didn't include an rf gain control, adding one can further improve the performance and versatility of the rig when attempting to copy weak signals in the presence of strong ones. An additional improvement is the complete elimination of popping. (This simple matter is explained in Ten-Tec's technical note TN2-525.) The modification consists of replacing the single potentiometer used for the audio gain with a dual concentric 10-k potentiometer. To install the new control, the front panel and the old control were removed. I soldered the wires from the old af control in the same locations on the front section of the new dual potentiometer by also inserting a 2.2-k carbon resistor in the center leg (the Argosy has so much gain that only about a third of the old potentiometer range was usable before). Then I connected the rear section as shown in **fig. 7**.

This change eliminated the switch function of the af-Power control. (This can be a good move because you won't need to worry about induced hum into the af sections any more.) The rig may be switched either from the front panel of the power supply or by a miniature switch added to the back panel. If desired, an AGC on/off switch can also be installed by breaking the connection between D9 and Q5 on the i-f/af board 80785 and wiring in another miniature switch. I chose not to implement this modification permanently because it wouldn't add anything to the performance.

Because concentric knobs weren't available, they were fabricated. The af knob was made by cutting off the top of a regular knob with a diameter of 0.8 inch. Additional filing brought the set screw within immediate proximity; little shaft length was available because of the spacer effect of the front panel. The rf knob was fabricated from a smaller diameter knob with the 0.25-inch hole filled with epoxy and redrilled for the inner shaft, which measured 0.125 inch in diameter. The entire modification took half an hour to complete.

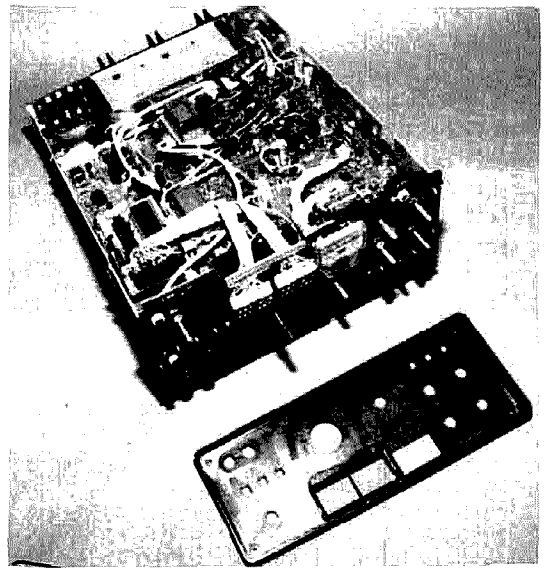


fig. 9. Installation of the digital counter in the Argosy and interconnections with the NSN 784 displays. Also shown is how the red window is secured to the front panel of the Argosy with epoxy through the dial opening.

PTO frequency stability

Another annoying problem related to the Argosy's performance was the frequency instability of the PTO, which caused small, sudden jumps of a few tens of hertz in the received or transmitted signals. This phenomenon seemed to be intermittent and showed up at random times with no apparent remedy. The problem, confirmed by other Argosy owners, may be present in similar rigs.

While the problem appeared mechanical at first, I concluded that it was caused by an intrinsic characteristic of a part in the design of the PTO. I was happy to find that Ten-Tec offers — on an exchange basis and for a reasonable fee of $25 — a new PTO that doesn't exhibit the problem. When a new PTO was obtained and installed, the problem disappeared once and for all.

The installation requires the removal of several wires and PC boards, but shouldn't take more than 15 minutes to complete. The performance of the new PTO is well worth the investment and allows the use of a precision digital counter which displays the transceiver's frequency with a resolution of 10 Hz, a considerable improve-

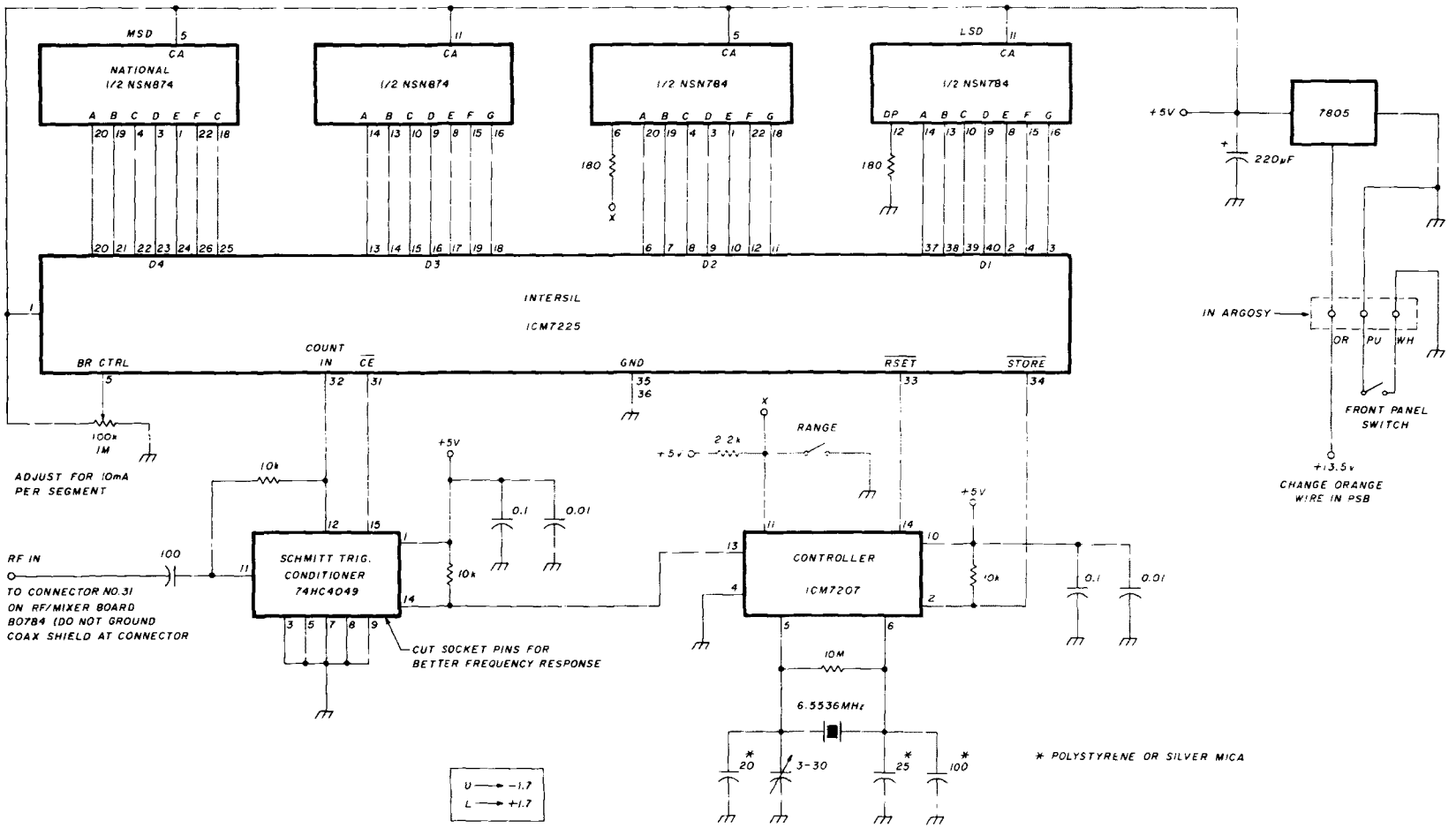


fig. 10. Diagram of the frequency counter as implemented in the author's Argosy. The entire board was wire wrapped in an afternoon on a 2 x 3.3 inch perforated board and installed in place of the calibrator board using its existing three holes.

ment over the ±2-kHz calibration provided by the Argosy's analog dial.

high-resolution digital display

The static digital display design in the modified Argosy uses three ICs and does away with the multiplexer noise so familiar to other designs. It allows the use of the front panel push button previously used to turn on the calibrator to further increase the resolution of the displayed frequency to 10 Hz, a feature I found extremely helpful.

The top-notch performance of the modified Argosy was finally complemented by the installation of a static (non-multiplexed) four-digit counter which displays the frequency of operation by reading the premixed PTO/-crystal oscillator's frequency. Reading the premixed injection presents the advantage of accounting for all possible shifts created in the crystal oscillators, which would have to be accounted for mathematically if reading only the PTO frequency. On the other hand, such a counter must be capable of operating at frequencies better than 20 MHz (as opposed to 5.5 MHz) if measuring the PTO. In addition, by wiring the calibrator push button from the front panel as shown, the four-digit numbers can be shifted to the left, allowing for 10-Hz resolution (an additional decimal point lights up to indicate that the magnifying feature is turned on). See **fig. 8**.

The design of the display was inspired by Intersil application notes. I had initially hoped that an LCD display would fit under the Argosy's front panel, but this was impossible because of the rather small dimension between the top of the bezel and the display opening. In addition, the 16-kHz back plane oscillator called for in this design was out of the question because of rf noise. The resulting design uses a static seven-segment LED design based on the ICM7225 LSIC from the same manufacturer.

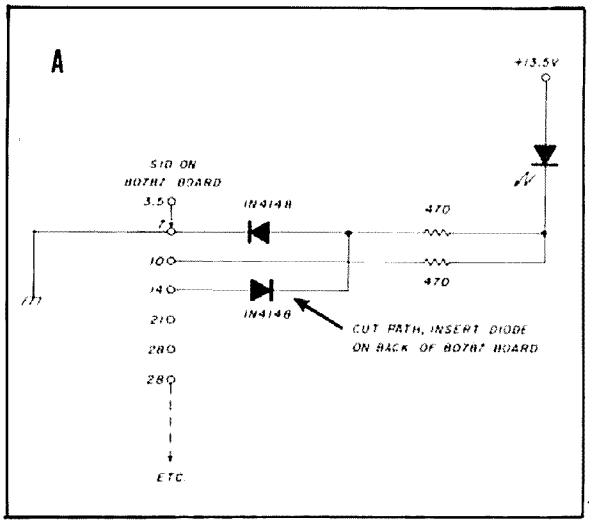
While many seven-segment LED displays were available, I chose the National NSN 784 common anode display because of depth restrictions between the front panel and the sub-panel. The 784 is built without pins and space-consuming extruded parts. Its design is basically a square PC board — with the LEDs fused into the surface of the board, allowing for minimal depth — which fits snugly between the two panels. With this approach, the flat cable used to connect the display with the small digital board located in the transceiver fits nicely through the narrow window previously used for the analog dial as shown in **fig. 9**.

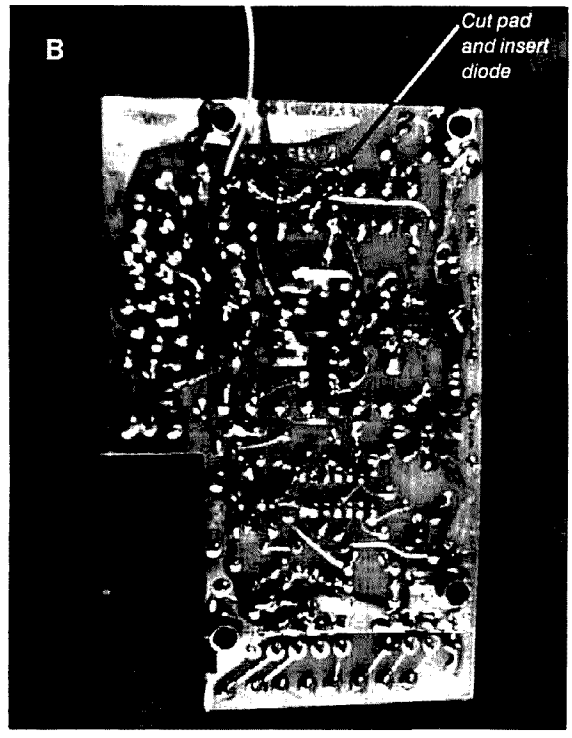
After this wiring was completed, the socket pins from the signal-conditioning IC 74HC4049 and from the controller/oscillator IC ICM7207 were shortened for better frequency response. If not short enough, pins on the 74HC4049 socket will limit the frequency response of the counter on the higher bands. As a matter of choice, the front panel switch (CAL) can be used either for turning on the display or for the magnifying feature. Both choices are shown in the schematic diagram in **fig. 10**. Only one function is possible without the addition of a separate switch.

The rf injection signal is fed to the conditioning IC 74HC4049 via a coaxial cable and a 100-pF capacitor from connector 31 located on the rf/mixer board 80784. When soldering at the connector, make sure not to ground the shield of the coaxial at this point or you'll defeat the turn-on function of the front-panel switch.

Adjusting the display is easy when using a high-resolution counter connected to the same pin on connector 31. Adjust the trimmer capacitor on the ICM7207 for an equal frequency readout between the counters, which should both be set in a high resolution mode. If a counter isn't available, reasonable results can be obtained by tuning to WWV. This counter will read only the center frequency of a station as offset by the BFO modifications performed earlier. If you're using the recommended settings, you'll have to add 1.7 kHz in lower sideband and subtract 1.7 kHz in upper sideband. Because no preset lines are available in this simple design. A feature activated from the band switch can tell the user when to add or subtract the offset number by lighting up either a plus sign or a minus sign inside the last digit of the display. This was accomplished by cutting a pad and installing a diode in its place instead on the S1D switch from board 80787 as shown in **figs. 11A, B,** and **C**.

Installation of the digital display in the Argosy proceeds as follows. Obtain two NSN 784 dual displays and grind off the rivets holding the windows to the back of the PC boards. Carefully remove the windows and plastic inserts so as not to touch the fused LEDs on board, now visible to the eye. Deposit a few drops of "Super Glue" or equivalent on the board in areas outside the segments and LED locations and key back the plastic insert without the window. Squeeze the assembly together until the two parts become a single assembly. This will prevent possible damage to the LEDs. Repeat the operation for the second display. Then solder a 16-wire flat cable to each of the boards as shown in **fig. 10**. Remove the front panel from the Argosy and remove the existing red window as well as all mechanical cables and fixtures associated with the analog display. File out the square opening in the front panel to accommodate the two new windows. Fit the two square windows snugly in the opening until they're flush with the outside surface of the front panel. Then deposit a small amount of epoxy around the perimeter of the two squares inside the front panel. In the process, make sure that the two windows are flush with the outside of the front panel; this is accomplished by laying the entire assembly on a flat surface. Only a small amount of epoxy should be deposited so that the PC board LED assemblies can fit





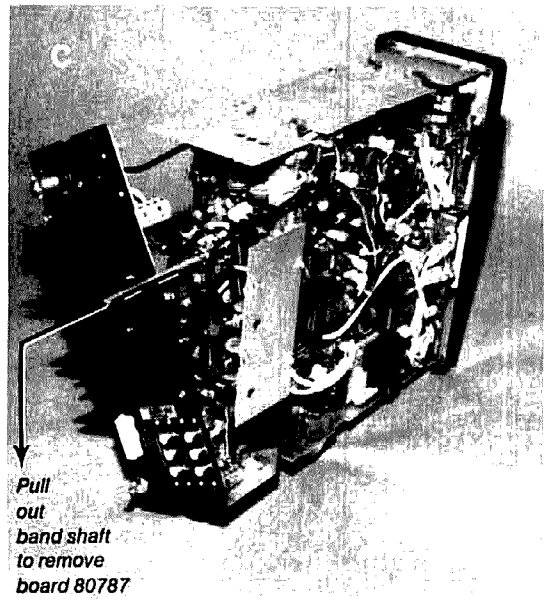


fig. 11. (A) Diagram of a sideband correction indicator (see text). (B) Implementation of the sideband correction mechanism on board 80787 requires cutting a pad and installing a simple silicon diode (1N914) in its place. (C) Removing Board 80787 from the transceiver requires the removal of the bandswitch shaft through the back.

back in their rivets when the job is complete. No epoxy should be allowed in the area under the top of the front panel because the cover must be free to slide back under the bezel.

To complete the job, wrap electrical tape around the metal cutout to prevent possible grounding. Place a weight on top of the front panel assembly and allow it to cure overnight. When curing is complete, reinstall the Argosy's front panel by pulling out the ALC indicator diode assembly and plugging it properly into the red LED from the front panel. Then carefully insert the two seven-segment LED boards in their keys, which are now attached to the front panel. Make sure that the translucent plastic diffuser squares are sandwiched between the red windows and the PC assemblies. Then pull things together until the digits are visible through the front windows. Push back the entire front panel assembly until the ALC LED fits back in its hole and screws can be attached. The flat cables can now be bent and properly wired to the digital board. To complete the job, make a 0.25-inch deep cutout in the top cover for the entire length of the display in order to prevent possible shorts. As an additional precaution, place electrical tape along the cutout.

conclusions

The modified Argosy (**fig. 12**) has been in operation for several months. Its performance has been a pleasure and well worth the efforts described in this article. Other possibilities exist for modifying the Argosy to increase its versatility. These include adding a separate PTO with memories; an rf splitter in the receiver input to allow for a separate receiver to be used in conjunction with the transceiver (this can be done by breaking the coaxial cable from connector 32 and introducing a Mini Circuits PSC2-1 rf splitter in the circuit with one of the outputs wired to one of the empty RCA connectors in

Table 3. Parts available from (A) suppliers and (B) author.

(A) Parts available from suppliers.

6.5536-MHz crystal, available from JAN CRYSTALS, P.O. Box 06017, Fort Myers, Florida 33906-6017 ($7.50 plus shipping).

XF-901 crystal, available from Spectrum International, P.O. Box 1084, Concord, Massachusetts 01742 ($6.25 plus shipping)

ICM7207 (don't order the "A" model), available from Advanced Computer Products, P.O. Box 17329, Irvine, California 92712-7329 ($7.50 plus shipping).

ICM7225 LSIC, available from Intersil authorized distributors nationwide (e.g., Arrow Electronics, Hamilton Avnet, Schweber Electronics, $12.50 plus shipping).

74HC4049, High-speed CMOS, 16-pin hex/buffer/converter (inverting), available from JAMECO Electronics, 1355 Shoreway Road, Belmont, California 94002. Digi-Key, P.O. Box 677, Thief River Falls, Minnesota 56701-9988. ($2.50 plus shipping).

(B) While quantities last, the following hard-to-find parts are available from the author.

Hewlett Packard 5082-2800, set of four matched high-voltage Schottky diodes ($15.50 plus $1.50 shipping).

Hewlett Packard 5082-3081, set of two matched high-current PIN diodes ($8.50 plus $1.00 shipping).

Dual concentric 10-k potentiometer for rf/af gain modification ($12.50 plus $1.50 shipping).

22 μF, 35 VDC Tantalum capacitor for AGC, $2.25 plus $1.00 shipping.

9.000-MHz i-f filter, 2.2 kHz (BW per specification), $45.00 plus $2.00 shipping.

Second 9.000-MHz i-f filter, 2.2 kHz (BW with blocking capacitors), $45.50 plus $2.00 shipping.

470 μH, low-resistance, high-Q inductor for audio LP modification ($6.00 plus $1.00 shipping).

Set of two 0.02-μF capacitors for audio low-pass filter, $0.50 plus $0.50 shipping.

Set of two NSN 784 LED seven-segment displays (four digits), $9.00 plus $1.50 shipping.

If combinations of the above items or the entire kit ($140.00) is required, include $3.00 for shipping (U.S. only, please).

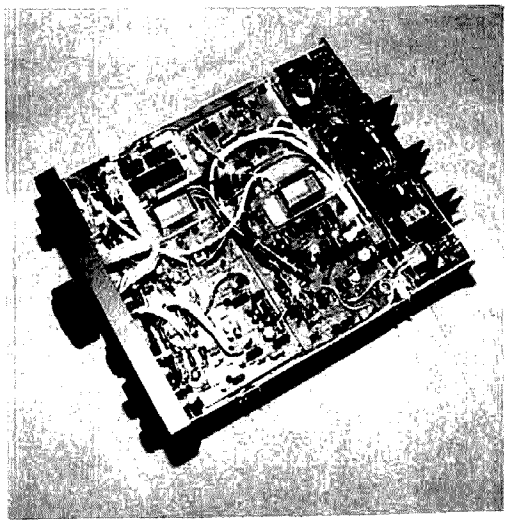


fig. 12. Completed unit.

the back of the transceiver); passband tuning in the i-f, and even a new BFO circuit. A list of parts available from the author and from various suppliers is shown in **table 3.**

acknowledgments

I would like to thank Tom Jorgensen, K0UBF, for donating his unmodified Argosy for the laboratory tests; Marc Denis, KD0QO, for being instrumental in performing them; Ed Wetherhold, W3NQN, for identifying the proper commercially available inductor and capacitor for the low-pass filter design; and Rick Whiting, W0TN, for his suggestions during the writing of this article.

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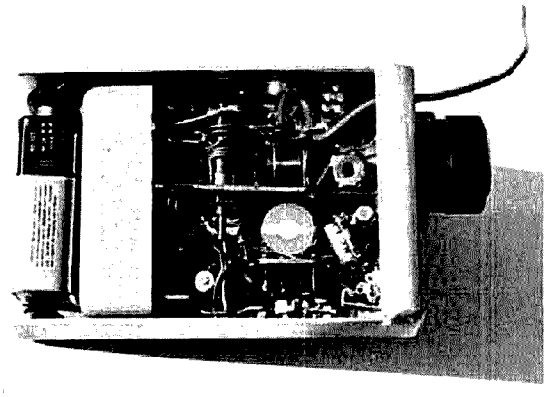
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ham radio

14-MHz superhet uses inexpensive ICs



build a pocket-portable SSB receiver

This article describes the design and construction of a very small, portable 20-meter single sideband receiver. Because some of the more demanding performance requirements can be relaxed in favor of small size and reduced component count, the project also represents an interesting design exercise in value engineering.

For example, reference 1 states that performance limitations are the principal drawbacks of integrated circuit mixers. Yet for a portable receiver in use in a remote location — say on a camping trip — small size and high gain may be more important than strong signal performance. Reduced complexity is also attractive to inexperienced builders; using ICs simplifies the design, since they can be considered as building blocks already designed by the manufacturer's specification sheets. The only real work, then, is selecting the best ICs and connecting them together.

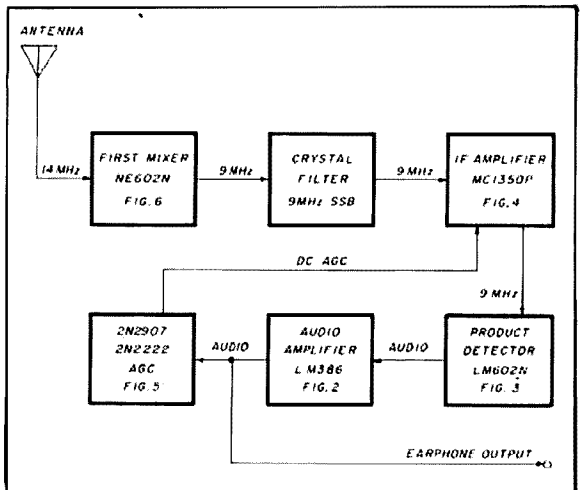


fig. 1. 20-meter receiver block diagram.

circuit description

Figure 1 shows that the receiver is a conventional single conversion superheterodyne design with a 9-MHz i-f. Major circuit blocks are represented by IC part numbers with the exception of the AGC. The following circuit theory of operation is presented in a sequence beginning at the output and proceeding toward the input, which is the recommended sequence for construction. Dividing a large project into several small tasks makes the work seem easier, and the completed output stages can be used as a test aid to align and troubleshoot preceding stages assembled later.

audio amplifier

Figure 2 shows the National Semiconductor LM386 Low-Voltage Audio Power Amplifier, which is used to drive an earphone. The main advantage of this component is low battery drain and operation from a 9-volt

Cliff Klinert, WB6BIH, 1126 Division Street, National City, California 92050

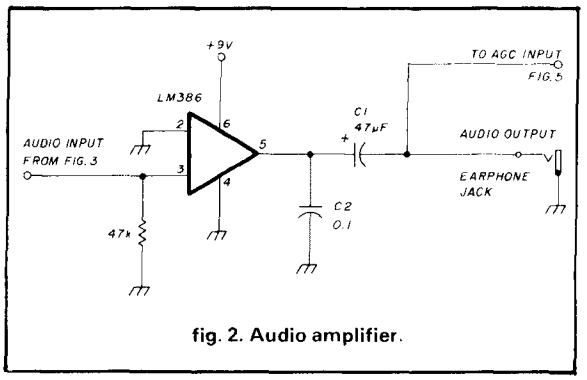


fig. 2. Audio amplifier.

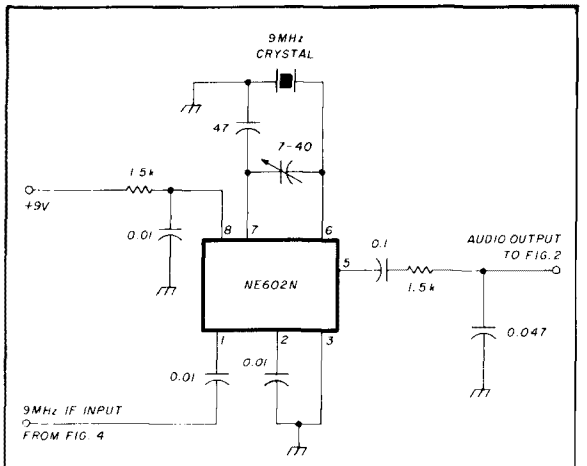


fig. 3. Product detector. Pins 4 and 5 are push-pull outputs, and either may be used.

battery. The 9-volt power source was considered essential for convenience in obtaining batteries while traveling.

Peripheral components increase circuit complexity, but provide an opportunity to shape the receiver audio frequency response. Signal frequencies below a few hundred Hz and above a few kHz should be attenuated by selecting audio coupling and bypass capacitor values carefully. Each capacitor (and associated resistance) provides a 6-dB per octave rolloff beyond the break point. The half-power break frequency is:

$$F = \frac{0.159}{RC}$$

where F is in Hz, R is in ohms, and C is in farads. For example, the output coupling capacitor, C1 of **fig. 2**, gives a low-frequency rolloff frequency calculated as 160 Hz. This is assuming that the LM386 output impedance is very low and the earphone offers the principal resistance in the circuit. I actually measured the earphone impedance to be 21 ohms at 1000 Hz when the calculations gave confusing results. The earphone is presumably 16 ohms, but it could have been 8 ohms. The following data was taken for selected capacitors using one-half voltage for the measurement point, which resulted in slightly lower frequencies (6 dB down):

330 µF 10 Hz
100 µF 70 Hz
47 µF 120 Hz

These results suggest that it may be better to substitute capacitors and sweep the frequencies with an audio oscillator than perform calculations when circuit resistance may be inconvenient to determine. In fact, components were selected experimentally, but the equation is helpful for awareness of circuit operation and to provide a starting point. A bypass capacitor (0.047 µF) in the product detector output provides a similar function for high frequencies, and the equation is the same.

C2, added to cure a high-frequency oscillation, may not be necessary in all cases. A large capacitor bypassing the audio power amplifier wastes power and may cause instability.

Tests showed that a 30-millivolt, 1000-Hz input signal provided a 600-millivolt output for a gain of 20. Current drain was 5 milliamperes at 9 volts.

product detector

The Signetics NE602 shown in **fig. 3** is used for the product detector as well as the first mixer. One of the more popular Gilbert Cell mixers, it also contains a voltage regulator and bias, an oscillator, and an rf amplifier. These features are implemented with an absolute minimum of external components. Pin 7, the oscillator emitter, can be used as a test point for checking oscillator operation. 350 millivolts of 9-MHz sine wave signal is a typical value. The capacitors connected to pin 7 must be adjusted for the correct operating carrier frequency to match the crystal filter. Power voltage for the NE602 is 6 volts, which is supplied to pin 8 through the 1.5k resistor. Application of a 400-microvolt 9-MHz i-f input signal to pin 1 gives a 150-millivolt peak-to-peak earphone output when connected to the audio amplifier **see fig. 2**. The current drain is an economical 2.5 milliamperes. More information on the NE602 is provided in the first mixer circuit description.

The resistor/capacitor network connected to pin 5 shapes the audio frequency response as described in the previous section. The 0.1 µF capacitor provides low frequency rolloff to complement C1 of **fig.1**. The 0.047 µF capacitor bypasses high frequencies.

i-f amplifier

A Motorola MC1350P television i-f amplifier was

used for the 9-MHz i-f amplifier **fig. 4.** This popular IC features high gain, excellent stability, and a very handy AGC input terminal. The MC1350P was intended for a 12-volt power supply, but works well on 9 volts with a current drain measured at 12 milliamperes. No particular effort was expended to wind T1 with a 14-turn primary (center tapped) and a 14-turn secondary on an FT 37-61 ferrite toroid core with No. 27 AWG wire. It seemed to resonate with about 15 pF of capacitance provided by a 9- to 20-pF variable, but it also works without the capacitor. Be sure to ground the AGC input, pin 5, for maximum gain while testing. Performance testing now shows that only 0.7 microvolts RMS 9-MHz input is required to produce a 100-millivolt peak-to-peak earphone audio output. The MC1350 gain was 50 dB with a 200 microvolt i-f output for the 0.7 microvolt input. Total receiver current drain at this point was 20 milliamperes.

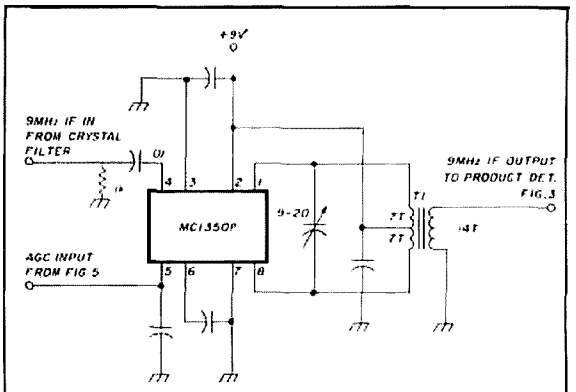


fig. 4. I-f amplifier. All capacitors are 0.01 disc except the 9-20 pF variable.

crystal filter

The crystal filter is the largest and potentially most expensive item. The price of a new crystal filter would make it impractical for so simple a project. I first considered the crystal filter as an optional or less critical item, but the obvious performance improvement made it a necessity. The filter and BFO crystal can be a junk box or surplus item. The 9-MHz filter used for this project was purchased for about $20 from a *ham radio* advertiser a few years ago, but swap meets and hamfests are also good sources. Be sure to get a BFO crystal with the filter, but if several filters are available inexpensively it may be better to break apart a filter to get an oscillator crystal. With luck it may be possible to tune the BFO to the right frequency with the oscillator capacitors and avoid the expense of ordering a crystal. At this point, a substantial effort may be required to sweep the filter with a signal generator while monitoring the receiver audio output and adjusting the BFO frequency for the desired frequency response.

automatic gain control

AGC is a truly optional feature that could be substituted with a potentiometer voltage divider connected to the MC1350 AGC input. Some experience using this method showed that the "two-knob" tuning method required was somewhat inconvenient considering the small controls and lack of panel space. Also, eliminating the manual gain control provided additional space for an AGC circuit.

The AGC characteristics of the MC1350 used were measured and tabulated as follows:

Pin 5 voltage	4.50 volts max gain
	4.77 volts min gain

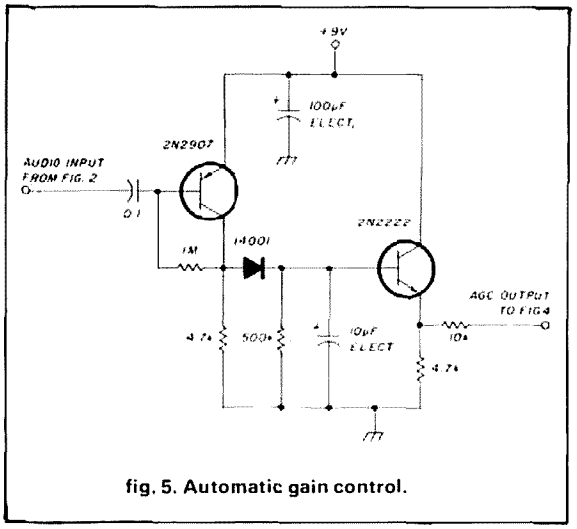


fig. 5. Automatic gain control.

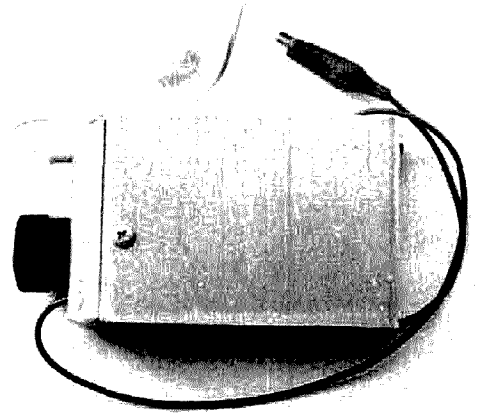


Photo A. The receiver is housed in a sheet aluminum enclosure. The aluminum was bent in a brake and the corners were TIG (Tungsten Inert Gas) welded. The clip lead is the antenna connector.

Pin 5 with a 10-k series resistor	4.00 volts max gain 6.30 volts min gain

Pin 5 is essentially a current sink. Adding a 10-k resistor in series makes a convenient voltage range for the AGC circuit to drive.

Figure 5 shows the AGC circuit, an unfortunate departure from the use of ICs. The Plessey SL621C would have provided an excellent AGC IC, but I felt that it was too sophisticated (and expensive) for this simple application. The 2N2907 amplifies the earphone audio, which is half-wave rectified by the 1N4001. The time constant circuit uses a 10-μF capacitor and a 500-k resistor. The 2N2222 is a dc amplifier that drives the MC1350. A 100-μF electrolytic capacitor was required to bypass the 9-volt power line to prevent a low-frequency oscillation.

The addition of this AGC circuit holds the earphone audio output to 100 millivolts peak-to-peak and consumes 2 milliamperes of current drain. Performance could be described as very aggressive but slightly slow on attack time. A slow attack time prevents the receiver from being muted for the RC time constant period when only a brief noise pulse is received. A full wave AGC rectifier would improve attack time but would also add to complexity. The circuit has more gain than is required to keep the audio ouput leveled, but I prefer this compression effect to eliminate the need for any other gain control.

mixer

The NE602N is used in the receiver first mixer (**fig. 6**). Two LC-tuned circuits are used in this circuit; L3 tunes the 14-MHz input and L1 is the 5-MHz local oscillator (LO) frequency control. L2 is a small rf choke added to restrict the L1 tuning range to 5.000 to 5.375 MHz. It's difficult to find a small vernier dial and variable capacitor, so a knob was attached to L1 for a tuning control. L1 was a 0.375-inch iron slug-tuned coil wound with 17 turns of No. 27 AWG wire. Many different L and C combinations could probably be used in this oscillator circuit. Pin 7 is the emitter and pin 6 is the base of the oscillator transistor. Reference 2 cautions against using too low a circuit *Q* and suggests connecting a 22-k resistor from pin 7 to ground if the oscillator won't start. Increasing the 300-pF capacitor values in **fig. 6** reduced the oscillator output, which was measured at pin 7 to be about 300 millivolts peak-to-peak. L3 is a 0.250-inch iron slug tuned form with 14 turns of No. 27 AWG wire. The input coupling link used five turns of hook-up wire. Don't try to use capacitor coupling for the NE602N input without a parallel choke or coil winding to ground. Low-frequency noise ("hum") will be coupled to the input amplifier, modulating all signals received.

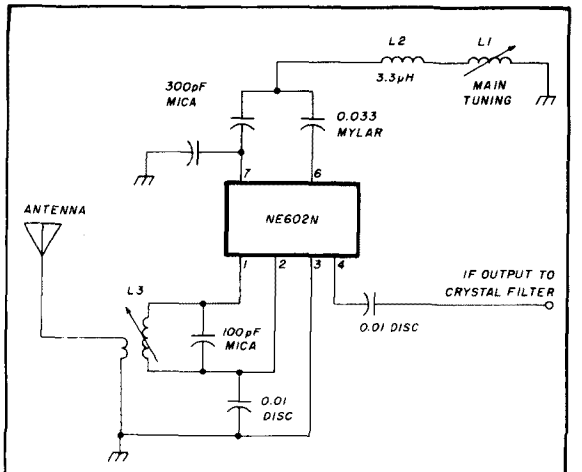


fig. 6. First mixer. The value of the 300-pF capacitors must be adjusted for alignment.

construction notes

I started out soldering components together by their leads as sort of a breadboard without the board to test the circuits. It was soon obvious that the most compact construction technique is to cut the leads as short as possible and solder them together on a double-clad unetched circuit board. The board serves as a ground plane as well as a structural mount. This mounting method is really quite strong, considering the light weight of the parts. A dab of epoxy would really add to strength, but it hasn't been necessary. Most of the parts were junk box selections so only general descriptions of the parts required will be given. This is especially the case for the tuning control, L1. The tuning slug screw must have a friction lock to prevent backlash. Since wear is a problem, select the largest screw size slug available and lubricate it with automotive wheel bearing grease. In this case, I found a short spacer with a tapped hole that fit on the tuning slug screw and locked it with a jam nut the same size. A knob can fit easily over the outside diameter of the spacer. The obvious disadvantage is lack of calibration markings, so it's important to limit the L1 tuning range to reduce confusion.

It may be desirable to enclose L1 and other oscillator components with circuit board since flexing the external cabinet will cause frequency changes. The double-clad printed circuit board is quite rigid when soldered together to make a box. After all the components were assembled and tested on the circuit board, I made a small aluminum box to fit around them. It's probably better to plan ahead and buy a small box, but at the beginning I wasn't sure what the final size or shape would be. (See **Photo A.**)

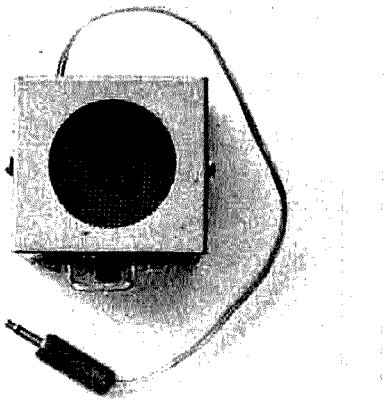


Photo B. This small box contains an amplifier and a speaker. The amplifier is the same as shown in fig. 2. No volume control is required when used with the receiver.

Soldering the components together does pose some problems, however. As can be seen in the photographs, it's very difficult to see some of the components. Repairs seem impossible. Don't solder to the crystal can near any of its solder seals; doing so may cause a vacuum leak, spraying molten solder on the quartz.

The electronic aspects of construction require use of an oscilloscope with a high impedance (x10) probe, an audio oscillator, and an rf signal generator. A frequency counter is handy but not essential. The x10 probe allows monitoring the oscillators with a minimal amount of frequency shift due to loading by the probe. I connect my frequency counter to the oscilloscope channel 1 output for checking the oscillator frequency. This is especially handy for setting up the BFO when the exact frequency is known, but the BFO frequency is subject to adjustment later "by ear" when listening to SSB signals. The first mixer oscillator can be adjusted with a receiver or signal generator, and will also require considerable adjustment to obtain the right frequency range.

conclusion

This receiver has enough gain to be used with almost any metal object as an antenna. I used a clip lead as an antenna connector. It's handy to carry a convenient length of insulated flexible wire for those times when there's a chance to rig a makeshift antenna in a tree or bush. Twenty meters provides good listening most times of the day or night, with some DX in the early morning.

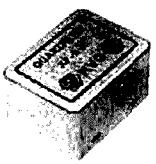
This article presented a few applications of some simple but versatile integrated circuits. My intent was to present a few circuits that might inspire the reader to develop new applications. For example, I mounted an LM386 with a battery and speaker in a small box to carry with the receiver. (See **Photo B**.) Future projects for the NE602 might include a tunable shortwave converter for an a-m car radio, or an aircraft band receiver with an MC1350 i-f amplifier.

references

1. Gary A. Breed, "Mixers: Making the Right Choice," *rf design*, August, 1986.
2. "Applying the Oscillator of the NE602 in Low-Power Mixer Applications," Signetics Applications Note No. AN1982, October, 1985. (Signetics Corporation, 811 East Arques Avenue, Sunnyvale, California 94088-3409.)

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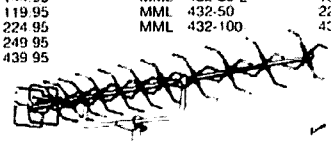
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A simple, low-cost homebrew project

20-meter hf superhet

One of the more diminutive boxing champions of the 1970s was described this way: "pound for pound, the hardest hitter in the world." I like to think that the same can be said of this receiver. Based on ideas put forth by Doug DeMaw and Wes Hayward,[1] it's strictly a lightweight, but it delivers a respectable punch. Its simplicity is accomplished through the use of a single i-f amplifier/detector IC, the Ferranti ZN414.

Because of its sharp selectivity, this superhet (**fig. 1, Photo A**) represents a vast improvement over direct-conversion receivers — though with fewer than 60 parts, it's hardly more complex. I built the prototype to tune about 125 kHz of the 20-meter band, but the receiver can easily be modified to cover 30, 40, or 80 meters.

circuit description

After passing through a fixed-tuned preselector, incoming signals at 14 MHz are admitted directly to gate 1 of the dual-gate MOSFET mixer. Mixer output at the intermediate frequency is passed through a tuned circuit at the primary of T1, a trifilar-wound transformer.* The transformer output supplied to the crystal filter consists of two signals of opposite phase. The effect is that the parallel capacitances within the two crystals are cancelled, leaving an extremely high-*Q* series-resonant circuit. This circuit configuration is known as a half-lattice filter. The shape of the filter's response curve is affected somewhat by terminating resistor R1. R1 is adjusted until the filter response is fairly flat across the top of the curve. The overall width of the curve is determined by the difference in the crystals' frequencies (see **fig. 2**).

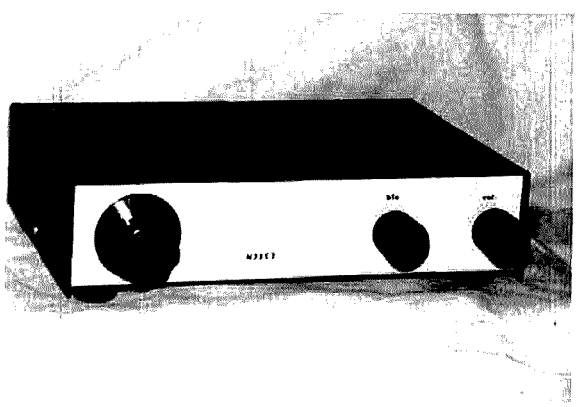


Photo A. The receiver cabinet measures 1 $^{15}/_{16}$ x 8 ¼ x 6 $^{1}/_{8}$ inch. (49 x 210 x 156mm). Main tuning knob is on the left. Lettering was accomplished with dry label transfers, and the panel was sprayed with clear acrylic.

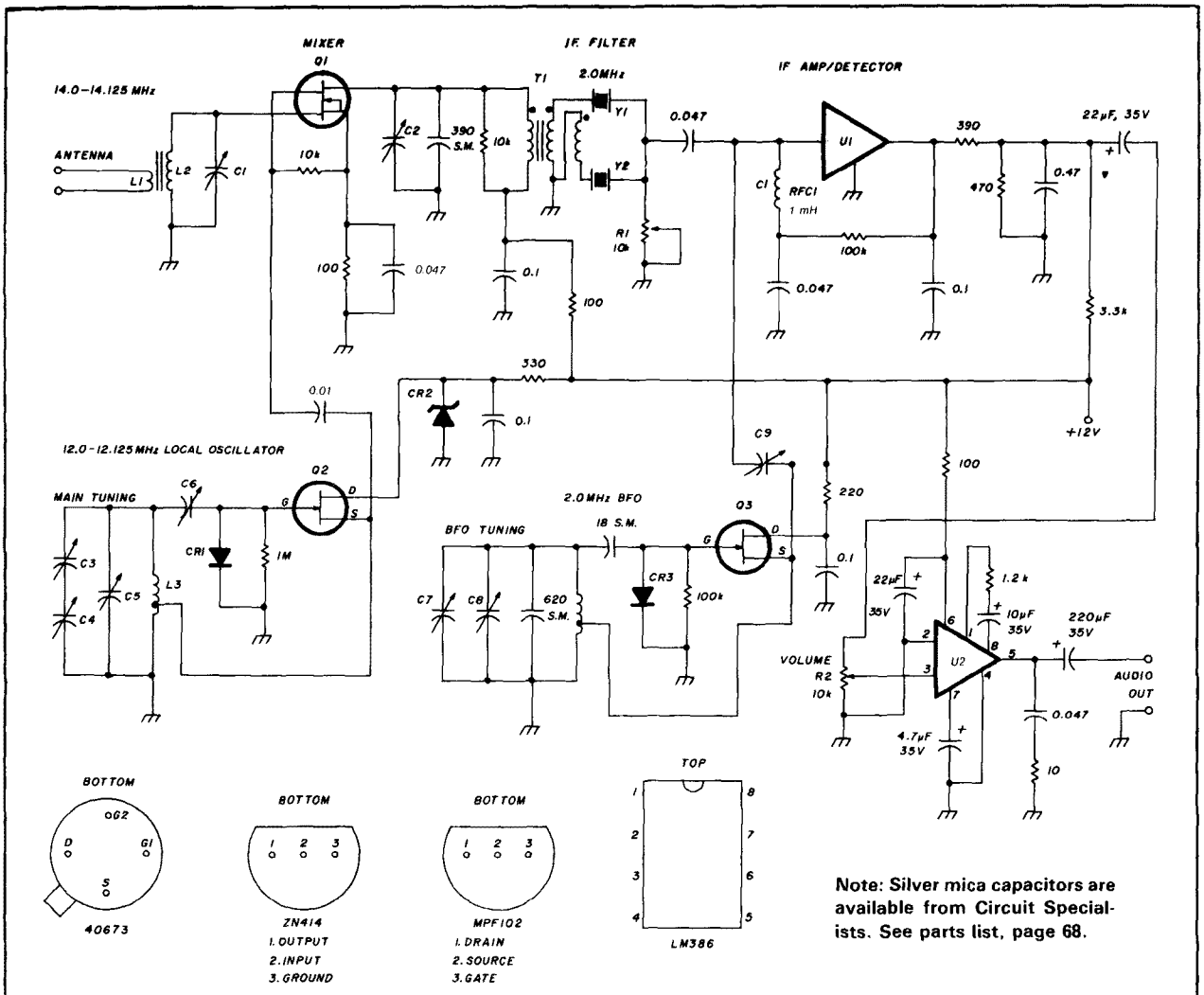
I used crystals with a frequency separation of about 600 Hz. With different crystal spacing the filter's bandwidth will change. If you wish to use this type of filter you can order two close-tolerance crystals for the desired frequencies. You can also buy three or four inexpensive crystals for about the same price; I used this less expensive approach and obtained three 2-MHz crystals for $2.50 each. When I checked the crystals (this can be done using a breadboarded oscillator with a frequency counter or receiver), I found a pair with spacing that was close enough, but not too close — in other words, just right!

There is an alternative for the experimenter who wants to to simplify this design even further. Acceptable performance can be obtained using only one crystal (**fig. 3**). Trimmer capacitor Cx is adjusted for the sharpest response. This filter has an asymmetrical response curve, with a fairly sharp peak, and one side quite broad. I tried this circuit and found that rejection of the adjacent sideband was fairly good, with only the stronger signals appearing on the wrong side of zero-beat. There's a sharp notch in which the adjacent sideband seems to be totally rejected; it appears that to bring out the best in this simple filter one should include it in a receiver using an audio bandpass filter peaked to the frequency where the image notch appears.**

*To make a trifilar winding, twist three separate pieces of enameled wire approximately 8 turns per inch. Wind the twisted wires through the toroid for 12 turns. I use an ohmmeter to identify the ends.

**If you happen to acquire two crystals of exactly the same frequency, you can use them to construct a ladder filter. I didn't try this approach, but details can be found in Reference 1, pages 217-218.

By Eric Bodner, N3ECZ, 64 South Hazle Street, Hazleton, Pennsylvania 18201



Note: Silver mica capacitors are available from Circuit Specialists. See parts list, page 68.

fig. 1. Schematic diagram. Judicious use of discrete devices and hardware results in a miniature receiver that's easy to tame and operate.

The filtered signal at the intermediate frequency is coupled to the input of the ZN414 i-f amplifier. The ZN414 consists of a very high gain rf amplifier followed by an AM detector. The device has three leads and comes in an ordinary TO-18 transistor package. This IC is hot enough to pick up the least bit of stray rf from any nearby (or distant) AM broadcast station, amplify it, detect it, and pass it on loud and clear. In fact, the chip makes a complete TRF receiver for any frequency up to 3 MHz with the addition of only a few external components. However, in the present application the chip should be restricted to amplifying and detecting only a narrow slice of the rf spectrum! Therefore shielded cable is a must for the connections to the ZN414 input from the i-f filter and the BFO.

Audio from the output of the ZN414 is sufficient to drive an LM386 amplifier IC, which in turn provides plenty of volume for headphones or a small speaker.

The local oscillator and BFO are identical except for the tuned circuits. NPO ceramic, polystyrene, or silver mica are the preferred types for the frequency-determining capacitors. C6 is kept at a minimal value to avoid oscillator pulling. Originally a 33 pF fixed-value capacitor was used at C6 and it seemed that every CW station on 20 meters was guilty of sending out a chirpy signal! With the smaller value capacitor in place, loading of the oscillator tank circuit is reduced and no trace of chirp is detectable.

I included a front-panel tuning control for the BFO, but this is optional. If you don't need the versatility

Parts list, fig. 1:

C1, C2, C5, C8	15-150 pF trimmer capacitor (Radio Shack 272-1339)
C3	15-90 pF air-variable with built-in vernier drive ($2.25 from BCD Electro)
C4	5-60 pF trimmer capacitor (Radio Shack 272-1340)
C6, C9	3-10 pF trimmer capacitor (Radio Shack 272-1338)
C7	1.4-13 pF air-variable ($3.63 from BCD Electro)
CR1, CR2	1N914 (Radio Shack 276-1122)
CR3	6.2-volt Zener (Radio Shack 276-561)
L1	2 turns around ground end of L2
L2	22 turns on a T-50-6 (approximately 1.9 μH)
L3	24 turns on a T-50-6 (approximately 2.3 μH) tap at 6 turns from ground end
L4	40 turns on a T-68-2 (approximately 9.1 μH); tap at 10 turns from ground end tap
Q1	40673 dual-gate MOSFET (available from Circuit Specialists Inc.)
Q2, Q3	MPF102 FET (Radio Shack 276-2062)
R1	10-k trimpot (Radio Shack 271-218)
R2	10-k pot (Radio Shack 271-215)
RFC1	1-mH choke (Circuit Specialists 43LS103)
T1	12 turns trifilar on an FT-37-61 (approximately 8 μH). Note: all toroids available from Circuit Specialists, Inc.
U1	Ferranti ZN414 a-m sub-system IC ($2.00 at Circuit Specialists Inc.) Also available from MHz Electronics, Inc., 3802 North 27th Avenue, Phoenix, Arizona 85017; OC Electronics, P.O. Box 3203, Scottsdale, Arizona 85257; Electronic Emporium, Inc., 4909 East McDowell, Phoenix, Arizona 85008; and Semiconductors Surplus, 2822 North 32nd Street, Unit 1, Phoenix, Arizona 85008.
U2	LM386 audio amp IC (Radio Shack 276-1731)
Y1, Y2	2.0 MHz crystals in HC6/U holder ($2.50 from BCD Electro). Y1 and Y2 can be resonant at any frequency between 1.7 and 2.5 MHz. At an i-f below 1.7 MHz, interference is likely from broadcast stations. Above 2.5 MHz it's likely that the ZN414 IC won't have sufficient gain

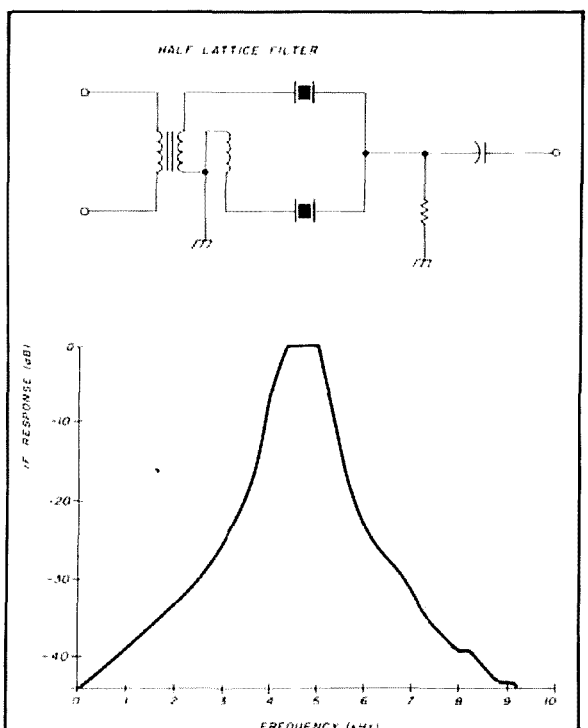


fig. 2. I-f stage response using half-lattice filter.

of a tunable BFO (for tuning either USB or LSB), then a PC-mounted trimmer capacitor may be substituted for the panel-mounted air-variable.

Coupling capacitor C9, which limits the BFO injection, should be adjusted for best results. In this circuit it's necessary to leak only a very low level of BFO energy for CW/SSB detection; the reason is that the BFO carrier is injected before the i-f amplifier. This differs from the usual method in which there is a separate mixer (product detector) and a fairly high-level BFO signal. I can justify the method used in this receiver only by saying that it allows for a substantial reduction in complexity (thanks to the circuitry built into the ZN414) and that I find it works quite well.

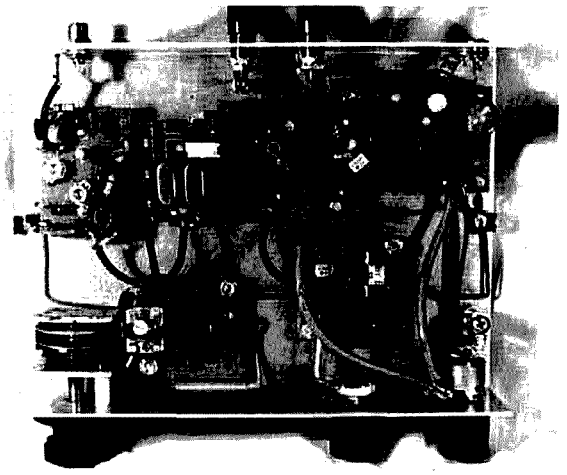


Photo B. Each stage is assembled on a separate board. Layout follows the schematic, with the mixer at the upper left and audio amplifier to the right.

construction notes

You can select your own scheme for wiring and assembly, bearing in mind the necessity for shielded cable at the input to the ZN414. The method I prefer involves assembling each stage on a separate circuit board so that construction, testing, and troubleshooting are limited to simple circuits containing about ten components each (see **Photo B**). It's much safer making the journey step-by-step, rather than attempting one giant leap!

Radio Shack circuit boards 276-024 are just the right size; extra holes are drilled through the large copper pads wherever they're needed. A sewing-machine needle makes a satisfactory miniature drill bit.

I've found that push-in terminals are practically indispensable for making external board connections such as ground, +12 volts, input, and output. These are available from Circuit Specialists, Inc. (No. 1495-PK). The circuit boards rest on aluminum spacers (Radio Shack No. 64-3024) and are fastened to the cabinet with machine screws. Solder lugs (Radio Shack No. 64-3029) are grounded to the cabinet in a similar fashion.

Use RG/174-U miniature coax cable for connections

on the signal path. This includes the connections between the various stages and the connections from the antenna connector and the volume control. RG/174-U is available from parts stores that serve the TV repair trade.

The crystals can be mounted directly on the circuit board, with the pins used as soldering terminals. Be sure to keep the crystal filter input and output well isolated.

Circuit layout at this frequency is generally noncritical. Still, it's best to position the local oscillator and BFO circuitry close to the front panel, with short runs of stiff wire to the tuning capacitors. Care should be taken in mounting the capacitors and toroids so that they are mechanically rigid.

The Radio Shack 270-272 cabinet made an attractive and sturdy enclosure for this receiver. Mounted on the rear panel are binding posts for power supply connections, a phono jack for the antenna, and a miniature phone jack for headphones or speaker (**Photo C**). The front panel layout includes the tuning knobs for the local oscillator and BFO, as well as the volume control and the on/off switch (Radio Shack 271-215).

access to parts

Radio Shack parts numbers are included for many of the components used in this project. The rest of the parts can be obtained by mail order.[2,3] The tuning capacitor with built-in vernier and dial pointer, available from BCD Electro, rates as a "best buy." Its threaded holes on the front and bottom allow easy mounting. For the prototype of this receiver I used a 30-pF air-variable without a dial pointer but with a built-in reduction drive almost identical to that in the BCD Electro capacitor that had been liberated from a used CB rig. You may choose to forego the use of mechanical bandspread and instead use a small variable in parallel with the main tuning control.

The cost of this receiver is low. Using all new parts the receiver can be built for around $40, but of course with help from a respectable junkbox the cost can easily be reduced to almost zero.

alignment

Before applying power, check the resistance between ground and the +12-volt line; it should be in the vicinity of 200 ohms. With the set turned on and the antenna connected, C4 and C5 should be adjusted for the desired frequency coverage. Tune C8 to place the BFO at the i-f. C1, C2, and C9 are peaked for maximum volume. While a frequency counter and a signal generator would be helpful during the alignment process, perfect results can be obtained with nothing more than a general-coverage receiver and a bit of patience.

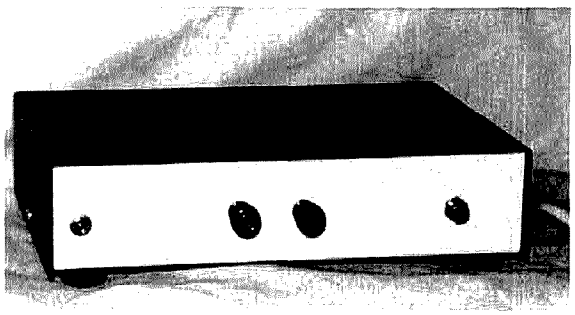


Photo C. Rear panel connections include (left to right) the audio output jack, binding posts for the DC supply, and phono jack for antenna input.

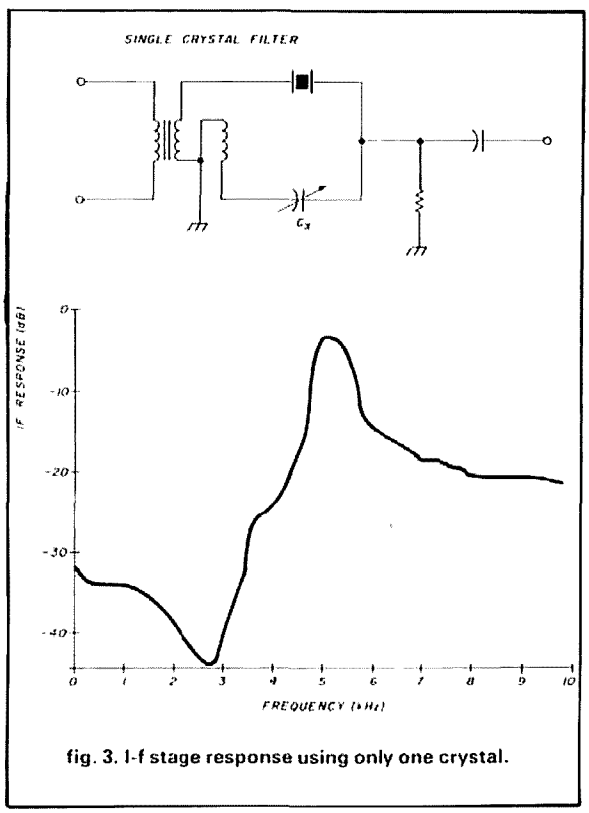


fig. 3. I-f stage response using only one crystal.

performance

Sensitivity was measured using an IFR-1200 Communications Service Monitor and a Motorola R2001-B Communications Systems Analyzer. An input signal of 0.88 microvolts rms was required for a

$$10dB\ \frac{S+N}{N}$$

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The i-f selectivity allows for good copy of both CW

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and SSB. Naturally the response isn't as sharp as that of a commercial receiver with eight poles of filtering, but for all but the most adverse band conditions, selectivity is more than adequate.

Stability is sufficient for casual CW work, but there's probably too much local oscillator drift for good results on RTTY or computerized CW. This is to be expected with an LC oscillator running as high as 12 MHz. Herein lies a tradeoff between performance and circuit complexity. A more stable (and more complex) local oscillator could be used at 12 MHz. (I've had good results with a series-tuned Colpitts VFO, but it also included buffer and doubler stages.) Alternatively, the simpler oscillator could be used down at 5 MHz for 40-meter reception, and stability would probably be very good. A crystal-controlled converter could then be used for 20-meter coverage.

other bands

I tried the receiver on 30 and 40 meters, with good results. Only one change was necessary. For 30-meter reception I simply retuned preselector C1. If C1 were replaced with a panel-mounted type, a 2-band receiver would result. For 40-meter reception, add a 220-pF capacitor across C1. The local oscillator can be tuned down to 9 MHz.

For operation on 80 meters, replace the antenna tuner and the local oscillator toroids; L1 will now have three turns; L2, 35 turns on a T-68-2 (approximately 7 μH); and L3, 30 turns on a T-68-2 (approximately 5.1 μH). Tapped seven turns from ground. Add a 220-pF capacitor across C1. A panel-mounted capacitor will be needed to peak the antenna preselector. A variation of 15 pF will allow for a preselector tuning range of approximately 100 kHz.

conclusion

The builder who has assembled a few simple projects will find this superhet easy to construct and a pleasure to operate. Perhaps this project will provide a bit of inspiration to some would-be homebrewer who has been discouraged by the supposed difficulty of building a complex receiver.

acknowledgments

Thanks to Tom Krohn, K3IJ, and Jim Debalko, N3BBX, for their advice and encouragement.

references

1. Wes Hayward, W7ZOI, and Doug DeMaw W1FB, *Solid State Design for the Radio Amateur*, ARRL, Newington, Connecticut, 1977.
2. Circuit Specialists, Inc., P.O. Box 3047, Scottsdale, Arizona 85257.
3. BCD Electro, P.O. Box 830119, Richardson, Texas 75083-0119.

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Bill Orr W6SAI

the super-cathode driven amplifier

Although the cathode-driven (grounded-grid) amplifier looks simple on paper, it is actually complex; what's more, some interesting variations of the basic circuit can provide unusual advantages to the user.

Consider **fig. 1**. This is a simplified grounded-grid amplifier circuit with a tuned cathode input. The drive signal (e_c) is link-coupled to the tuned circuit, but a pi-network or other arrangement could be used as well. The grid of the tube is at RF ground potential and the exciting signal is applied between cathode and (grid) ground.

Now look at **fig. 2**. All remains as before, except that the grid of the tube is tapped on the cathode tank so that a portion (e_g) of the total drive signal is applied in-phase to effectively oppose the cathode signal (e_c). This is termed a *super-cathode driven* amplifier. Drive power is thus increased and stage gain is decreased.

While it may appear ridiculous to design an amplifier that demands more than the minimum amount of driving power, this circuit may be used to advantage when it's necessary to absorb excess drive power from the exciter. A case in point: an amplifier is desired to run 1 kW PEP input and deliver about 650 watts output. It will be driven by the usual RF transceiver, with a power output of slightly over 100 watts. Aha! A single 3-500Z, running at 2500 volts and 400 mA, should do the job. A nice home project for the enthusiastic builder!

Looking at the 3-500Z data sheet shows that the idea is practical except for one fact. The 3-500Z requires only about 46 watts of drive power. What to do with the other 60 watts of drive power available from the exciter?

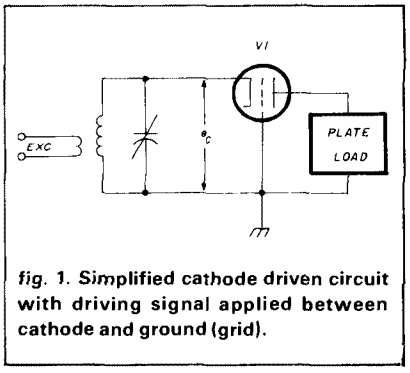


fig. 1. **Simplified cathode driven circuit with driving signal applied between cathode and ground (grid).**

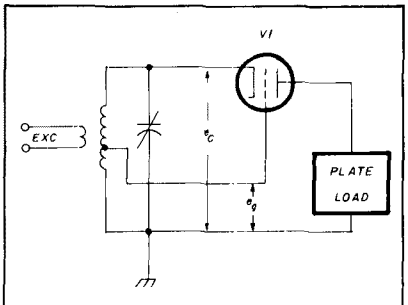


fig. 2. A portion of the drive signal (e_g) is applied in-phase to the tube. This opposed the larger cathode drive voltage, E_c. This is termed a super-cathode driven amplifier.

Some operators who have run into the problem of excess drive try to solve it by merely cranking down the mic gain. This may work if they don't get excited and run the gain up, and if the exciter is capable of proper ALC action at reduced input (many aren't).

Perhaps what's needed is inverse feedback in the amplifier to "soak up" excessive drive. This will help, but only marginally. A T-pad between exciter and amplifier? *Yes, but unless the pad* is switched in and out, it will remain in the circuit during reception . . . not an elegant solution.

super-cathode driving

This is a case where the super-cathode driven circuit comes into its own. Excess drive power is converted into "feedthrough" power and stage gain can be tailored to fit the available level of drive.

Hold it! Don't rush out and start converting your 3-500Z amplifier to a super-cathode driven mode. Removing the grid from ground destroys the inherent isolation of the stage and instability is likely to result. What's required is a 3-500Z with an extra grid to be used for RF shielding purposes! Luckily, such a tube has been available for decades: the 4-400A.

The 4-400A super-cathode driven tetrode amplifier

At one time the 4-400A was widely used by Amateurs in grounded-grid amplifier service. Merely tie both grids of the tube to ground and away you go. While the 4-400A wasn't designed for this class of service, it did work, provided the user kept a close eye on grid current . . . it was easy to exceed the No. 1 grid wattage dissipation limit of the 4-400A in classic grounded-grid service (see **table 1**).

The 3000-volt conditions listed in

1987 CALLBOOKS

147



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Table 1. 4-400A (Voice Conditions). Ratings apply to 4-250A, within plate dissipation rating of 4-250A.

DC Plate Voltage	2000	2500	3000	volts
Zero-Signal DC Plate Current*	60	65	70	mA
Single-Tone DC Plate Current	265	270	330	mA
Single-Tone DC Screen Current	55	55	55	mA
Single-Tone DC Grid Current	100	100	100	mA
Single-Tone Driving Power	38	39	40	watts
Driving Impedance	160	150	140	ohms
Load Impedance	3950	4500	5000	ohms
Plate Input Power	530	675	990	watts
Plate Output Power	325	435	600	watts

*Varies from tube to tube

table 1 are interesting. The simplified circuit is shown in **fig. 3**. Grid and screen terminals of the tube are grounded and the drive is applied to the filament circuit in the conventional manner. Power input is 1 kW PEP and drive power is about 40 watts – about the same as a 3-500Z.

But the big advantage of the 4-400A is that it has two grids to play with. Grid No. 1 can be used to vary the power gain of the tube and grid No. 2 provides excellent internal shielding.

In passing, it should be noted that when used in a super-cathode driven circuit, the drive signal serves as a screen power source. Screen-to-cathode voltage (e_{sg}) is supplied by the drive signal as shown in **fig. 4**. The control grid of the 4-400A is tapped to a point on the cathode circuit to provide the desired amount of "bucking" voltage (e_b) and the screen element is at RF ground potential. Resting plate current is low because screen voltage is low (zero) with no drive signal. Converted drive power is large, as is total grid drive requirement.

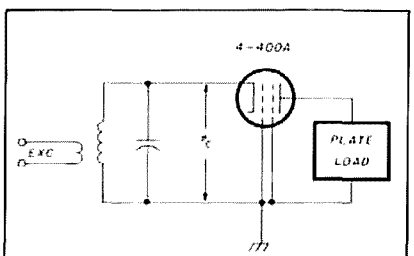


fig. 3. The 4-400A tetrode connected as a high-mu triode for cathode driven (grounded grid) service. Screen-to-cathode voltage (e_{sg}) is equal to e_c.

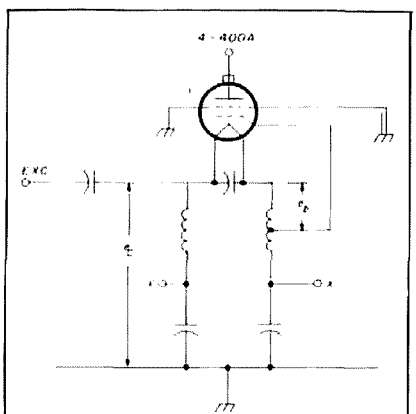


fig. 4. Drive signal (e_c) is developed between cathode and ground. Bucking signal (e_b) is developed between cathode and grid #1.

the K7BYQ-W7EPM tests

In 1959 two EIMAC engineers, K7BYQ and W7EPM, ran extensive tests on a super-cathode driven amplifier using a single 4-400A tube.[1] The results of their tests are listed in **table 2**. They had difficulty getting the test amplifier to work properly on the highest bands (20 and 10 meters) because ferrite-core filament chokes weren't available in those days. But the amplifier worked perfectly on the lower bands, and when a good filament choke was finally designed, performance was duplicated on the higher bands.

However, in 1959, hams weren't troubled by an excess of driving power,

Table 2. 4-400A (voice conditions), super-cathode drive.

DC Plate Voltage	3000 volts
Zero-Signal DC Plate Current*	30 mA
Single-Tone DC Plate Current	330 mA
Single-Tone DC Screen Current	60 mA
Single-Tone DC Grid Current	30 mA
Driving Impedance	250 ohms
Load Impedance	5000 ohms
Plate Input Power	990 watts
Plate Output Power	625 watts
Single-tone Drive Power	115 watts

*varies from tube to tube

the cathode-driven circuit hadn't yet become popular and the whole idea ended up in the dusty pages of history.

Perhaps the super-cathode driven amplifier is today's solution to a 1 kW PEP amplifier driven by a 100-watt exciter. The 1959 experiments showed that the ratio between screen and grid voltages for a 4-400A in this circuit was about 5:1. The peak screen driving signal was about 320 volts. The peak grid signal, then, is about one-fifth of this, or about 64 volts.

How to obtain this voltage ratio between grid and screen voltages? K7BYQ and W7EPM tapped the filament choke (**fig. 5**) directly. One side of the filament was grounded to complete the DC return. Screen current was monitored as shown.

Another scheme is to add a third winding, one having one-fifth the number of turns of the bifilar winding, to the choke. Thus, if the choke has 30 bifilar turns, six turns could be added to achieve the 5:1 voltage ratio (**fig. 6**). The ratio can easily be altered by changing the number of additional turns.

If the number of turns between grid No. 1 and the filament (cathode) is too small, the amplifier resembles a straightforward grounded-grid stage. Driving power will be a minimum and grid current will be very high. As the number of turns is increased, grid current drops and grid drive level rises. Too many turns will make the tube very hard to drive.[2]

The data in **table 2** is enough to get started. It will be interesting to see how this circuit works in conjunction with the solid-state exciters in common use today. Experimenters! Here's a chance to try something different.

For a reprint of the article describing K7BYQ's and W7EPM's work with cathode-driven amplifiers, send four first-class stamps or IRCs to me at EIMAC, 301 Industrial Way, San Carlos, California 94070. Request bulletin AS-33.

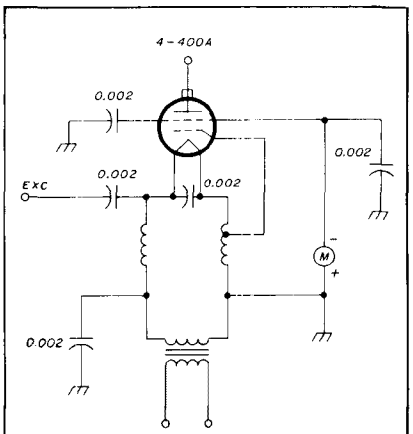


fig. 5. Original K7BYQ-W7EPM super-cathode driven circuit. Grid tap is one-fifth of the way from the filament end of the choke. Note same choke winding is grounded at filament transformer.

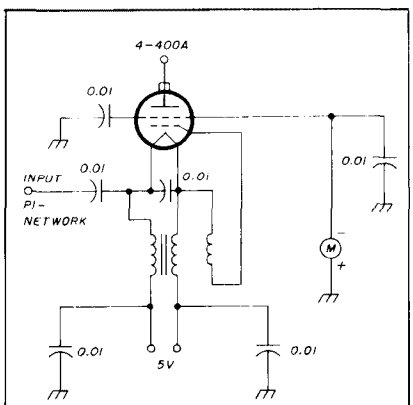


fig. 6. Auxilary winding can be used instead of tap on main choke winding. Meter measures screen grid current.

references

1. V.S. Campbell and W.S. Skeen, "Grounded Screen-grid Operation for Tetrodes," *QST*, November, 1959, pages 37-39.
2. William I. Orr, W6SAI, and William H. Sayer, WA6BAN, "Semi- and Super-Cathode Driven Amplifiers," *QST*, July, 1967, pages 33-38.

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PRACTICALLY SPEAKING ... Joe Carr K4IPV

from the mailbag

Although I usually devote this entire column to a single topic, I'd like to take this opportunity to respond to some of the questions readers have asked since this column began just over a year ago.*

parts sources

Several readers have asked where they can obtain parts for projects. If they're looking for standard parts, I refer them to ads in the pages of *ham radio* and other magazines. Sometimes parts needed for Amateur Radio projects can be found through local distributors, even though their primary market may be defined as consumer electronics service shops or computer hobbyists. Most semiconductor devices that you'll need are available from sources such as *Jim-Paks* or one of the consumer electronics replacement lines — for example, HEP, ECG, and NTR.

Radiokit** offers a 1500-watt transmatch kit that can be used for a variety of applications besides the stated purpose of antenna matching. Because it contains a 28-μH rotary inductor, turns counting dial parts, and a pair of 208-pF, 3500-volt variable capacitors, it can be adapted for use in homebrewed power amplifiers. Judging from my mail about high-power rf parts, I suspect that the age-old habit of building linear and Class-C power amplifiers is not yet dead — despite the large number of commercial amplifiers on the market.

Obtaining parts for power amplifiers can be aggravating. While every local ham shop used to carry such parts, they are now difficult to obtain. One chap complained bitterly about the difficulty in obtaining the high-voltage plate transformer for a linear project. The parts list called for a commercial transformer, but the manufacturer was used to dealing only with commercial and industrial clients. There was a minimum factory order of $500, and no means for dealing with clients who lacked bank or Dun & Bradstreet references. For high-voltage, high-power parts for dc power supplies, try Peter Dahl Company. They make transformers suitable for amateur projects — and will actually deal with you!

For a one-of-a-kind power amplifier project, of course, you can scour the hamfests and surplus shops. Unfortunately, while such parts were easily obtained from World War II surplus when I was a kid, those sources have since dried up. If you intend to write an article about the project for this or any other Amateur magazine, please don't use a part that was obtained at a hamfest, from a surplus house, or from under a buddy's workbench until you know for sure that the part is still available from other sources. For replication of your project by others, there has to be a reliable source of parts!

There are some novel ways to find high-voltage power supply parts. For example, I needed an oil-filled high-voltage capacitor for a project. Now, if you've watched the right TV shows, you've seen physicians use a clever device called a defibrillator to "jump-start" patients who try to die of certain kinds of heart attacks. The paddle electrodes are placed on the patients's chest, buttons are pressed and *BLAM!* — the patient's body lurches into the air a foot or so.

Because I'd worked as a biomedical engineer in a university hospital, I'd had some experience with defibrillators. The old standard defibrillators generated a damped capacitor discharge waveform from a 16-μF capacitor. Most of them consisted of a 7-kV dc power supply, a 16-μF capacitor, and a dpdt vacuum relay that transferred the capacitor terminals from the dc power supply to the patient electrodes (through a 100-μH choke). Typical defibrillators of a dozen years ago used a 16-μF/7500 wVdc or 16- μF/10,000 wVdc oil-filled capacitor that rarely went bad. Those capacitors make ideal filter capacitors in ham power supplies. The transformers are useless for this application because of their current rating, which is too low.

The vacuum relays (spdt in older machines, dpdt in later models) are useful for rf purposes. I used a relay from a scrap defibrillator to switch coil taps in a loaded vertical antenna many years ago.

OK, so how do you go about locating old defibrillators? They're not just lying around in surplus parts stores, are they? Yes, they are. I also saw one, an old American Optical defibrillator,

*Readers are welcome to contact me at P.O. Box 1099, Falls Church, Virginia 22041.

**Radiokit, P.O. Box 973, Pelham, NH 03076.

at the Gaithersburg Hamfest just a year ago. You might try calling the biomedical engineer at a nearby hospital or checking with firms that deal in used hospital or veterinary equipment. Defibrillators are high-risk items for hospitals; malfunctions cost lives and result in malpractice suits. As a result, outdated equipment is often taken out of service while the relay and capacitor are still useful to hams. Look for models by American Optical (16-μF/7500 wVdc), Hewlett-Packard (16-μF/10,000 wVdc or very old Air-Shields (25 μF/5000 wVdc). Don't pay a lot for these machines. When dickering over the price, remember that they're too risky for the smart hospital engineer to put back into service — and for the same reason, less useful for replacement parts than might be expected. Other sources of parts include rf diathermy equipment, industrial inductive heating equipment, magnetizing equipment, and so forth.

amplifier tuning problem

One reader complained that his commercial linear amplifier worked poorly, resulted in less than encouraging signal reports and (here's the clue) always tuned to the same spot on the loading dial. Furthermore, according to signal reports, the output was horribly distorted on 20 meters. My first thought, confirmed in conversations with two friends who work in radio transmitter repair, was that although something might be wrong with the amplifier, the most likely problem was in the antenna or transmission line.

In my own experience, the only time my transmitter (and later linear) loaded to the same point on the loading control dial was when the antenna (or coax) was either open or shorted. This problem was sometimes masked by the fact that the receiver still worked correctly. The 20-meter distortion was merely more pronounced than distortion on other bands. One explanation of the distortion would be improper impedance matching between the anode of the power tube and the antenna. Another possible explanation — one consistent with my experience — would be that the shield on the coax was open, leaving the reception intact but obliterating the transmitter or linear tuning! If you're experiencing that kind of problem, check out the antenna before messing with the transmitter.

In most commercial equipment, "loading problems" are unlikely if the correct load impedance is presented to the output. Of course, they're possible . . . but they're so unlikely that it's best to look for other causes first. The correct way to diagnose the problem, in my opinion, is to use a low-cost piece of test equipment that every Amateur ought to have: a 50-ohm dummy load. If the amplifier loads into the dummy, then I suspect the antenna.

Of course, it should go without saying that the Amateur should check the antenna, transmission line, couplers, low-pass TVI filters, and switches or relays in the line. In some cases, dc resistance checks of switches and relays is not too useful at rf. While a "bad" reading is to be believed, a "good" reading might be a false indication. If the dc resistance checks reveal no obvious fault, try checking the parts by either substitution or removal. In a future column, we'll deal with antenna instrumentation and troubleshooting.

coax velocity factor

The velocity factor (VF) of a transmission line is a decimal fraction describing the percentage of the speed of light (C) at which signals propagate in the line. For example, when we say that ordinary coaxial cable has a vf of 0.66, we are saying that the signal travels in the line at 66 percent the speed of light. When we want to cut a length of coax to a specific electrical length, we calculate the physical length required for that frequency and then multiply by the VF. A common example is the half-wavelength transmission line. The impedance of a load (such as an antenna) is repeated every half-wavelength back down the line. If we want low-cost antenna measuring instruments to be accurate, then we should use a connecting transmission



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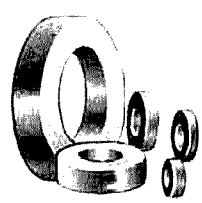
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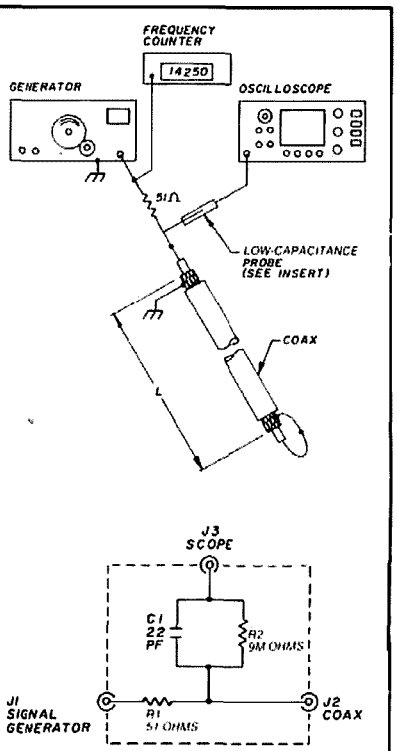


fig. 1. Simple, accurate method of determining velocity factor of coax.

line that is an integer multiple of half-wavelength. To find the correct physical length, we would use the half wavelength formula multiplied by the VF of the coax:

$$L = \frac{492}{f} \times VF$$

where:

L is the length in feet

f is the frequency in megahertz

VF is the coax velocity factor

When we work through an example, we find that a 20-meter half-wavelength (492/14) is 35 feet. If we use regular coax (VF = 0.66), then we find a correct physical length to achieve an electrical half-wavelength is (35 x 0.66), or 23.1 feet.

Standard Amateur literature (including mine) usually cites the following velocity factors for common coax:

Regular Polyethylene: 0.66
Polyfoam: 0.80
Teflon: 0.70

We take these figures as gospel true, only to find problems on the workbench. I discovered this problem years ago in a marine radio shop (the old-fashioned kind, where a "frequency meter" was a war surplus BC-221) where we used coaxial cable as a calibrator for depth sounders. Those devices send an ultrasonic pulse to the bottom and then time the return echo. We used a precisely cut roll of coax to time the signal to exactly 100 feet.

But the physical length I was ordered to cut didn't make sense if the coax VF was 0.80. The shop owner informed me (and he turned out to be right!) that most coax only approximates the specified velocity factor. He showed me a method of measuring VF in any given sample of coax (see **fig. 1**). The procedure is as follows:

1. Cut a convenient length of coax (20 to 100 feet). Measure its physical length as accurately as you can.
2. Connect the coax to a signal generator and either an oscilloscope or rf voltmeter as shown in **fig. 1**. (The junction box shown in the inset is useful.) Short-circuit the distant end of the coax.
3. Calculate the approximate half-wavelength frequency from

$$f = \frac{492 \cdot VF}{L}$$

using the "standard wisdom" velocity factor given above.

4. Adjust the signal generator output to maximum and get a decent deflection on the oscilloscope. Rock the frequency back and forth to find the frequency at which the trace on the oscilloscope drops to its lowest point. This frequency (which can be read on the optional frequency counter) is the electrical half-wavelength frequency. It is nulled because the short circuit at the end is a zero-ohms impedance, and is reflected exactly to the feedpoint at the half-wavelength frequency.
5. Calculate the actual VF from $(f \times L)/492$.

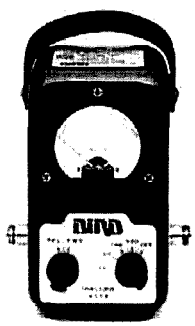
For example, I bought a 50-foot roll of RG-58 coax of unknown ancestry at a hamfest (it measured 51 feet). Because it was foam coax, the standard-wisdom VF was 0.80. Thus, a 51-foot length was an electrical half-wavelength on a frequency of (492 x 0.80)/51 ft), or 7.72 MHz. When I measured it using the method shown in **fig. 1**, the "short-circuit frequency" that indicated half-wavelength was not 7.72 MHz, but 7.13. Thus, the actual VF was

$$\frac{7.13 \cdot 51}{492}, \text{ or } 0.74.$$

While the actual velocity factor isn't terribly important for most Amateur antenna work, it can be important in measurements of antenna parameters, or in cases (like my marine shop employer's) where the coax is used in another measurement system.

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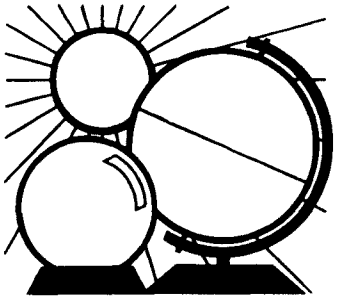
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DX FORECASTER

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sunspot minimum update

You can expect current DX propagation conditions to last for about a year after we pass the sunspot number (SSN) minimum; it will take about that long before much change can be noted. But when is the minimum? Have we reached it yet? Let's review the data so far.

Since our last look (in April, 1985) the SSN has shown more flattening, sinking to a value of 16 at the end of 1985. Even the increased solar activity in February and March, 1986 raised the SSN only moderately. In June, a new low in observed sunspots for this 11-year cycle was reached, but more activity occurred in the following months. Was June the minimum? At that point, the present sunspot cycle was ten years old (see "DX Forecaster," October 1985, page 105).

The new cycle (No. 22) region data analysis shows that an active region or two may have been seen for a day or so in April, May, and July of 1985. From September, 1985 on they were more numerous, were seen by more observatories, and lasted long enough for magnetograms to be obtained for polarity determination. In January, 1986, those of reverse polarity were seen. February's flares and mammoth geomagnetic disturbance were caused by a *Cycle 22* sunspot.

The SSN minimum should occur about a year after the first spots of the new cycle are observed. So is that September, 1986, or January, 1987? Your guess is as good as mine.

second corroboration

Now let's look at the solar flux (10.7-cm emissions) data. The solar flux minimum should be near, but probably not coincident with, the SSN minimum; it should occur within approximately one month of the SSN minimum in data corrected for distance. In determining the solar flux minimum (at the sun), scientists correct for the distance, producing a different minimum in their data than solar flux data from WWV.*

Prior cycles solar flux minima were recorded in June, 1954, July, 1964, and June, 1976; notice all were in the summer. There were 10.08 and 10.87 years between these minima, respectively.

In 1986, the solar flux increased in February and March — as did the SSN — then decreased steadily until June. If the solar flux minimum for Cycle 21 occurred in June, 1986, with a monthly average value of 67.5 and seven daily values of 66 in a row from June 24 to 30, the duration of the interval between the 1976 minimum and the 1986 minimum, then, would have been 11.05 years.

One can expect a daily value lower than 66 because the other minimum daily values were 63.0, 64.7, and a 66, which occurred during Cycle 20, lower in maximum amplitude than Cycle 21. Therefore we may expect a lower daily minimum value during the coming months.

Could a lower daily minimum value occur next summer? If it did, this would mean that the duration of the solar flux cycle would be at least 11.93 years. Not likely! Scientists at the Space Environment Center at Boulder have concluded that considering all the variables, the SSN minimum will probably occur during the first half of 1987. This forecast gives a Cycle 21 duration of 10.5 to 11 years, which is a reasonable length. Time will tell how all this analysis of data will turn out.

* A change in the earth-sun distance (e.g., between summer and winter) equates to a change in the included solid angle and consequently the amount of intercepted solar flux. Scientists normalize to one A.U. (astronomical unit – earth-sun distance) and WWV measures the data directly. – Ed.

last-minute forecast

The lower frequency bands, 30 to 160 meters, are expected to be excellent during the first two weeks of November. Summer and fall thunderstorm noise should be behind us, leaving quiet conditions with low QRN and also few geomagnetic disturbances. The maximum usable frequencies are low enough that 80 meters is near optimum for east-west and northern DXing at night. The higher bands may have some good openings during the second and third weeks. Expect some transequatorial openings in the evenings. They could be enhanced by sunspot activity and/or geomagnetic disturbances. There is the possibility of low signal strengths for a few days at a time, during both day and night, over the next few months.

The Taurids meteor showers will occur from October 26 to November 22, with a maximum count of ten per hour from the 3rd through the 10th of November. Lunar perigee is on the 4th, and a full moon appears on the 24th.

band-by-band summary

Ten and twelve meters, the highest day-only DX bands, are nearest the MUF for southern hemisphere paths. They will be open most days when the solar flux is above 75 during the 7- to 10-hour period centered around local

WESTERN USA

GMT	PST	N	NE	E	SE	S	SW	W	NW
0000	4:00	30	40	20	12	12	10	10	20
0100	5:00	20	40	20	20*	12	10	10	20
0200	6:00	30	40	20	12	20	12	12	20
0300	7:00	40	40	20	20	15	15	15	30*
0400	8:00	40	40	30	20	20	20	20	40
0800	9:00	40	40	30	20	20	20	20	40
0600	10:00	40	40	30	20	20	20	20	40
0700	11:00	40	40	30	30	20	20	20	40
0800	12:00	40	40	30	30	20	20	30	40
0000	1:00	40	40	30	30	20	30	30	40
1000	2:00	40	40	30	30	30	30	30	40
1100	3:00	40	40	30	30	40	30	30	40
1200	4:00	40	40	30	30	30	30	30	40
1300	5:00	40	30	15	15	30	30	30	40
1400	6:00	40	20	12	12	20	30	30	40
1500	7:00	40	20	12	12	15	20	20	40
1000	8:00	40	20	10	10	15	20	20	40
1700	9:00	40	20	10	10	15	20	30	40
1000	10:00	40	30	10	10	12	15	20	40
1000	11:00	40	40	10	10	12	12	20	40
2000	12:00	40	40	12	10	12	12	15	20
2100	1:00	40	40	12	10	12	12*	15	20
2200	2:00	40	40	15	10	12	10	12	20
2300	3:00	40	40	20	10	12	10	12	20

MID USA

MST	N	NE	E	SE	S	SW	W	NW	CST
5:00	30	40	20	12	12	10	10	40	6:00
6:00	30	40	20	15	15	12	12	40	7:00
7:00	40	40	30	20	20	15	15	40	6:00
6:00	40	40	30	20	20	20	20	40	9:00
9:00	40	40	30	20	20	20	20	40	10:00
10:00	40	40	30	20	20	20	20	40	11:00
11:00	40	40	30	30	20	20	20	40	12:00
12:00	40	40	30	30	20	30	30	40	1:00
1:00	40	40	30	30	20	30	30	40	2:00
2:00	40	40	30	30	30	30	30	40	3:00
3:00	40	40	30	30	30	30	30	40	4:00
4:00	40	40	15	30	30	30	30	40	5:00
5:00	40	30	12	20	30	30	30	40	6:00
6:00	40	20	12	15	20	20	20	40	7:00
7:00	40	20	10	12	15	20	20	40	8:00
8:00	40	20	10	12	15	20	30	40	9:00
9:00	40	20	10	10	15	20	30	40	10:00
10:00	40	20	10	10	12	15*	20	40	11:00
11:00	40	20	10	10	12	15	20	40	12:00
12:00	40	30	10	10	12	12	15	30	1:00
1:00	40	40	12	10	12	12	15	20	2:00
2:00	40	40	12	10	12	10	12	20	3:00
3:00	40	40	15	12	12	10	12	30	4:00
4:00	40	40	20	12	12	10	10	40	5:00

EASTERN USA

EST	N	NE	E	SE	S	SW	W	NW
7:00	40	40	20	20	15	15	15	40
8:00	40	40	30*	20	20	20	20	40
9:00	40	40	30	20	20	20	20	40
10:00	40	40	30	20	20	20	20	40
11:00	40	40	30	20	20	30	30	40
12:00	40	40	30	20	20	30	30	40
1:00	40	40	30	30	20	30	30	40
2:00	40	40	30	30	30	30	30	40
3:00	40	40	30	30	30	30	30	40
4:00	40	40	30	30	30	30	30	40
5:00	40	30	15	20	30	30	30	40
6:00	30	20	12	20	30	30	30	40
7:00	30	20	10	15	20	20	20	40
8:00	40	20	10	12	20	20	20	40
9:00	40	20	10	12	15	20	20	40
10:00	40	20	10	12	15	20	20	40
11:00	40	20	10	10	12	20	20	40
12:00	40	20	10	10	12	20	20	40
1:00	40	20	10	10	12	15	20	40
2:00	40	30	10	10	12	12	20	40
3:00	40	40	12	10	12	12	15	30
4:00	40	40	15	10	12	12*	12	20
5:00	40	40	20	12	12	10	12	20
6:00	40	40	20	15	12	15	12	30

NOVEMBER	WESTERN USA	MID USA	EASTERN USA
ASIA FAR EAST	3:00	4:00	5:00
EUROPE	4:00	5:00	6:00
S. AFRICA	6:00	7:00	8:00
S. AMERICA	7:00	8:00	9:00
CARIBBEAN S. AMERICA			9:00
ANTARCTICA	8:00	9:00	10:00
NEW ZEALAND	10:00	11:00	12:00
OCEANIA AUSTRALIA	11:00	12:00	1:00
JAPAN	1:00	2:00	3:00

noon. These bands open on paths toward the east and close toward the west. The paths may be as long as 2400 miles in single-hop length, and occasionally twice as long during evening transequatorial openings.

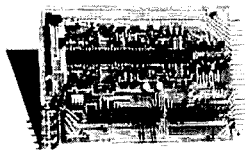
Fifteen meters, a day-only DX band open most of each day, has lower signal strengths and greater multipath variability than 10 and 12 meters. It will be best when the MUF is just resting above this band, until it drops below it — a transition period that occurs right after sunrise and just before sunset. Transequatorial openings will occur, with distances similar to 10 and 12 meters.

Twenty, thirty, and forty meters are both daytime and nighttime DX bands. Twenty is the maximum usable band for DX in the northern directions during daytime. In combination with 30 meters, it provides nighttime southern paths for the day-only bands. Forty meters becomes the main over-the-pole DX daytime band, with some hours covered by 30 meters.

Eighty and one-sixty meters, the night-only DX bands, exhibit short-skip propagation during daylight hours, then lengthen at dusk. These bands follow the darkness path, opening to the east just before local sunset, swinging more to the north-south near midnight and ending up in the Pacific areas for a few hours before dawn. On some nights 80 meters, with its high signal strengths, will be the best band to use. One-sixty is also expected to provide favorable conditions. Please remember the DX windows of 3790 to 3800 and 1825 to 1830.

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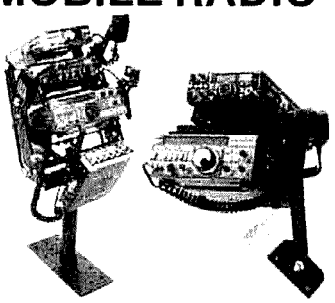
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VHF/UHF WORLD

Joe Reisert W1JR

broadband amplifiers in receiver design

Electronic technology is roaring forward. Sometimes that's bad, but generally speaking, it's good for Amateur Radio. Prices decrease, devices get smaller, and sometimes, performance improves! However, when technology changes, we often must learn new tricks or change the way we do things if we're to take advantage of the latest developments. This is a small price we pay.

Such is the case of small signal rf amplifiers. Only 20 years ago the Nuvistor was a premium device used in many of the low-noise VHF and UHF converters. Then came the bipolar transistor, which was followed by the JFET, MOSFET, and the GaAs FET. Each one of these devices caused us to change the way we designed our gear. We went from multiple power supplies with hundreds of volts and filament voltages to single low-voltage supplies, often below 5 volts.

At the same time, the circuitry has changed dramatically. Tubes often required large boxes and complex neutralization circuitry. New techniques such as grounded-gate JFETs or dual-gate MOSFETs solved some of the neutralization problems. Special feedback and loading techniques alleviated most of the remaining solid-state stability problems. Finally, broadband, unconditionally stable circuitry arrived. This, too, required a change in thinking because filtering was not contained within the circuit itself.

Traditionally November marks the appearance of the annual *ham radio* issue spotlighting receiver technology. The subject of receivers has been a part of *VHF/UHF World* from the start. Several areas of receiver design — including rf amplifiers, mixers, local oscillators, filtering, and high-dynamic range — have been covered.[1-5] In keeping with that custom, this month's column will be devoted to receiver-type rf amplifiers and more specifically, broadband amplifiers and how they can be applied to new or existing receivers.

rf amplifier design trends

In the past, most Amateur gear used narrowband tuned amplifiers (**fig. 1A**). This was good because the selectivity decreased out-of-band signals as well as the image frequency. These same tuning circuits performed the impedance matching required for the device in use. Dynamic range was usually enhanced, but input circuit losses often increased the noise figure.

Today's design trend is to use "gain blocks" that cover a wide band of frequencies. Usually these broadband circuits are rf amplifiers optimized for similar input/output impedances such as 50 ohms or 75 ohms, with the latter used mostly in the CATV business. Because these amplifiers often cover many octaves of frequency (each octave is a 2:1 frequency range), they require additional external filtering (**fig. 1B**). This often consists of a bandpass filter at the input and possibly one on the output of the amplifier. If desired, these filters may be located external to the amplifier and mounted in a shielded box following modular construction techniques described in reference 1.

The number of broadband small-signal rf amplifiers presently available is large, surely in the hundreds. Each has a unique set of specifications that include rf bandwidth, gain, impedance match, noise figure, and output power handling capability. The principal active devices used are bipolar transistors and GaAs FETs. The most common frequency range is 1 to 500 MHz. Many bipolar broadband amplifiers are specified to work up through 6 GHz. Some GaAs FET devices work well up into the millimeter-wave region.

amplifier evolution

To gain an appreciation for the changes in technology, it may be worthwhile to review some of the techniques used to design these broadband amplifiers. While it is theoretically possible to design broadband compensation networks, they are complex and may be lossy. The principal breakthrough for broadbanding is in the proper application of wideband negative feedback.

The earliest practical broadband circuits that I am aware of used discrete transistors with shunt and series feedback (see **fig. 2A**). The founders of Avantek discovered that a broadband impedance match could be realized if a proper high-frequency transistor was used with certain combinations of shunt R_F and series R_E feedback. The patented equation is as follows:

$$R_O^2 = R_F \cdot R_E$$

where R_O is the input/output impedance, R_F is the shunt and R_E is the series feedback (including the internal impedance of the base-to-emitter junction of the transistor) per **fig. 2A**.

The ratio of these two resistors de-

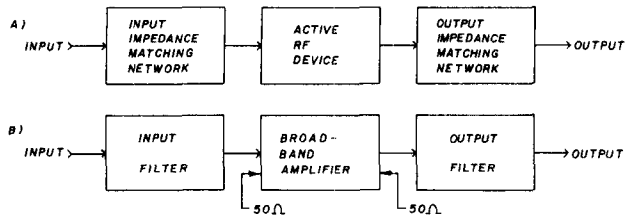


fig. 1. Typical rf amplifier configurations: (A) use of impedance matching transformers in a narrow-band design; (B) use of wideband amplifiers with matched input/output impedances and external filtering.

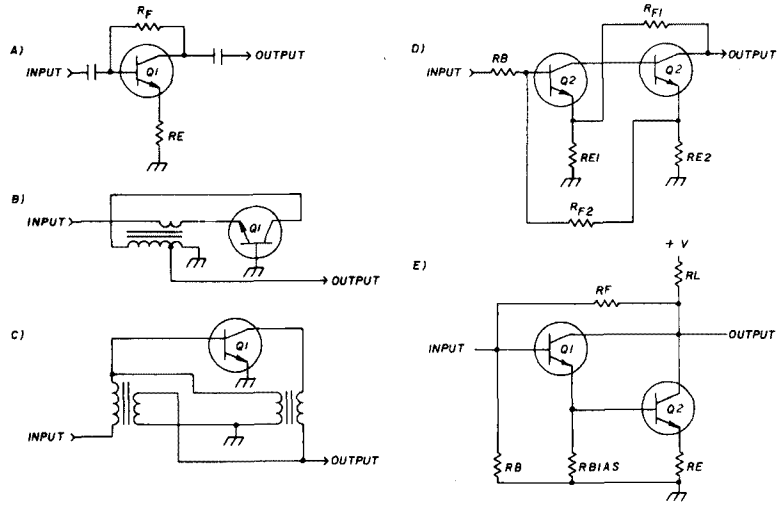


fig. 2. Typical rf representation of various feedback techniques. In some cases the dc portion of the circuit has been deleted for clarity. (A) shows series R_e and shunt R_f negative feedback; (B), lossless feedback employing transformer matching[3]; (C), feedback employing bybrid couplers[7], and (D), two-stage series and shunt feedback.[10] (E) shows Darlington connection, a version of fig. 2A[10,11] R_b and R_{bias} are bias resistors to establish the correct DC circuit parameters.

termines the gain, which can be approximated using the following equation:

$$G = 20 \log \frac{R_F - R_E}{R_O + R_E}$$

where G is gain in dB and R_F, R_E and R_O are as described above. For example, if R_F is 500 and R_E is 5 ohms, the amplifier will be impedance matched to R_O, 50 ohms, and will have a gain of approximately 19 dB.

The higher the value of the series feedback resistor, R_E the lower the shunt feedback resistor, R_F must be to maintain the match and vice-versa. Low ratios of R_F/R_E yield lower gain, higher noise figures, and lower output power compression points. High ratios of R_F/R_E yield greater gain, lower noise figures, and higher output power compression points. Although this feedback technique is patented, the patent is near expiration.[6] Reference 2 showed a typical application of this circuit with proper dc biasing.

"Lossless feedback" techniques have also been patented.[7,8] **Figures 2B** and **2C** are some typical rf circuits. The circuit shown in **fig. 2C** is often referred to as *directional coupler feedback*.

The principal advantage of lossless feedback is that there is very little impact on noise figure or output power due to the feedback circuitry. However, these circuits are more difficult to implement over a wide frequency range because of the complexity of the feedback transformer. Reference 2 showed an application of lossless feedback with a typical circuit.

Most of the early broadband rf amplifiers used discrete components and were somewhat limited in performance. In premium designs, "tweaker" inductors and/or capacitors were often added to tune out any circuit or component mismatches. Cost was moderate to high.

In the early 1970s, hybrid techniques using thin or thick film resistors, inductors, and capacitors — as well as chip capacitors and transistors — were popular. These units exhibited improved performance, but were still moderate to high in price.

Meanwhile, integrated circuit manufacturers were also trying to design broadband amplifiers. However, the techniques and materials in the substrates often limited the performance to an upper frequency limit of 100 MHz.

More recently, bipolar IC designs have moved up in frequency. These are referred to as *silicon microwave monolithic ICs* (MMICs). The Signetics NE5202 will operate up to about 600 MHz, while the NEC MM765, MM766, and μPC1656C and their related families will operate to just over 1 GHz.[9] These IC circuits all use a two-stage shunt and series feedback circuit that I found described in reference 10 (see **fig. 2D**).

High-performance, high-frequency (i.e., up through 6 GHz), single-stage, bipolar MMIC designs have now arrived in packages the size of an ordinary microwave transistor![11] The circuitry used is related to that shown in **fig. 2A**, except that some manufacturers use a Darlington connection of two transistors, as in **fig. 2E**. Improved transistors, substrate materials, and photo-lithography now allow very small geometries with greater control of parasitics. This yields more uniform

performance and a higher frequency of operation at a much lower cost.

Finally, GaAs FET MMICs are being manufactured that will significantly increase the state-of-the-art frequency limits with lower noise figures than bipolar MMICs. The less expensive lower-gain devices use circuits similar to **fig. 2A**. These GaAs FET amplifiers show great promise, although they are more costly than bipolar devices at the present time.

applications

So where does that leave us? The answer is that if we want to take advantage of the low-cost, small size, and higher performance MMICs, we have to again adapt to the recent changes in technology.

MMICs are going to be a way of life. They're small in size, require only a minimal number of external components, are broadband, easy to bias (most require only a single supply voltage and possibly one external resistor), unconditionally stable, and offer a good impedance match over many octaves of frequency. Furthermore, they're easily replaceable as new and higher performance units become available.

Let's examine some typical applications. MMICs have nominal noise figures of 3 to 7 dB (see **tables 1** and 2). Therefore, their noise figures are more than adequate for the first stage in moderate performance or low-cost receiver front ends, especially for local contacts. Many of the MMICs available also have moderate output power (10 to 20 milliwatts or +10 to +13 dBm), making them ideal for second-stage or post-amplifiers in receivers or i-fs.

Figure 3A shows a typical hookup for a "generic" MMIC. Some typical circuit board layouts with their amplification characteristics were described in reference 12, so they will not be repeated here.

MMICs are usable for remote or second-stage preamplifiers when tower-mounted gear is used. They're ideal as broadband amplifiers for use ahead of scanners or receivers that cover a wide range of frequencies, especially if they're outside the Amateur bands.

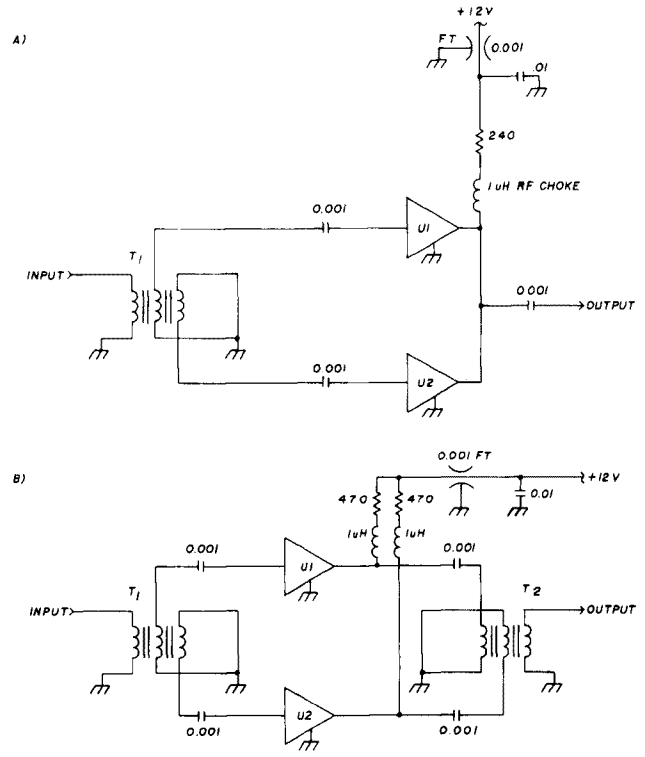


fig. 3. Typical applications using MMICs. (A) is a broadrand amplifier. C1 and C2 are dc blocking capacitors, C_b is a bypass capacitor, FT is a feedthrough capacitor, L is an rf choke and R is the biasing resistor chosen depending on the manufacturers design. Typical values are 100 to 1000 ohms. (B) shows an amplifier with two isolated outputs. Either a hybrid in-phase splitter or a Wilkinson power divider can be used.[5] Each output will be approximately 3 dB below the input. (C) is a typical applications of an LO amplifier with extra filtering.

Many Amateurs are using MMICs for LO (local oscillator) amplifiers, particularly where two outputs are desired for transverter applications, per reference 5 (see **fig. 3B**). A further application in LO service is where there is insufficient power to drive a DBM (double-balanced mixer). This is particularly true above 1 GHz, where some of the multipliers used have low output or low multiplier efficiency (see **fig. 3C**).

In a quick experiment, I hooked up two MMICs in a push-push configuration to see how they would perform as a balanced multiplier. I was quite pleased with the results. **Figure 4A** shows a preliminary circuit which could be further optimized or possibly use higher-power MMICs.

The output of this balanced doubler is free of spurious responses and works well through 500 MHz. This could undoubtedly be extended upwards in frequency with a better transformer design. Gain is moderate, 5 to 8 dB, especially when the input level is below 1 milliwatt. External filtering is desired, but it can be very minimal because the undesired outputs are typically at least 20 to 25 dB below the desired output signal.

If the MMIC outputs are hooked up in push-pull instead of push-push (**fig. 4B**), they should perform as a moderate gain tripler. I haven't tried this configuration, but it should be feasible.

Table 1. Some of these reasonably-priced commercial bipolar MMICs may be of interest to Amateurs.

Part Number	Supplier	Frequency Range GHz	Typical 0.5 GHz gain(dB)	Noise Figure (dB)at 500 MHz	Output Power dBm	Voltage/ current (mA)	Price each $
MAR-1	Minicircuits	DC-1	12	5.5	0	7/20	1.50
MM765	NEC	0.01-1.1	19	5.5	10	10/43	7.20
MM766	NEC	0.01-1.3	18	5.5	5	5/20	7.20
MSA-0104	Avantek	0.05-2.5	17	5	1.5	5/17	2.75
MSA-0204	Avantek	0.05-2.5	12	6	4.0	5/25	2.90
MSA-0304	Avantek	0.05-2.5	12	6	10	5/35	3.00
MSA-0404	Avantek	0.05-3	8	6	12	5/50	3.25
MSA-0185	Avantek	0.05-2.5	17	5	1.5	5/17	3.15
MSA-0285	Avantek	0.05-2.5	12	6	4.0	5/25	3.25
MSA-0385	Avantek	0.05-2.5	12.5	6	10	5/35	3.35
MSA-0485	Avantek	0.05-3.8	8	6	12	5/50	3.55
MSA-0835	Avantek	0.5-6	28	3	12.5	8.5/36	11.25
MWA0204	Motorola	0.01-3	12	6	4	5/25	2.65
NE5205	Signetics	DC-0.5	19	6	4	6/24	2.80
PC1656C	NEC	0.01-0.75	19	5.5	10	10/43	4.15

Table 2. The following reasonably-priced commercial GaAs FET MMICs may be of interest to Amateurs.

Part	CGY-40	NEPA 1001
Supplier	MSC/Siemens	NEC
Frequency range	0.5-4	0.5-3
Typical 0.5 GHz gain (dB)	9	11
Noise figure at 500 MHz	3	4
Output power (dBM)	17	9
Voltage /current (mA)	+5/60	5/60 −5/4
Price	$16.30	$23 @ 5M quantity

However, the harmonic output will require slightly more complex filtering because the number of output frequency components will increase.

suppliers and devices available

So far I've mentioned only general trends in circuit design and capabilities of hybrid circuits and MMICs. There are many suppliers of each. Some of the major suppliers of 50-ohm broadband hybrid rf circuits are Anzac, Avantek, Aydin Microwave, Hewlett Packard, MiniCircuits Labs, Motorola, Optimax Division of Alpha Industries, Q-bit, TRW, and Watkins Johnson. Suppliers of MMICs include Avantek, Minicircuits Labs, Motorola, MSC/Siemens, NEC, and Signetics.

Table 1 is a list of the most common low-cost bipolar MMICs that are available to Amateurs. **Table 2** lists some low-cost GaAs FET MMICs. Just as with transistors and FETs performance, the variations in frequency, gain, noise figure, and other parameters, are many. Hybrids and MMICs are presently available up to several watts of output from dc through the millimeter-wave range!

Probably the biggest thing for us to do as Amateurs is to become familiar with the devices that are available and use them wherever possible. In this regard, the GaAs FET MMICs are the most exciting because they hold the key to lower noise figures, higher frequencies of operation, and higher output power.

summary

Time and space have not permitted a long dissertation on the subject of hybrid and MMIC broadband rf amplifiers. Instead, only an overview has been presented. The references provided and data sheets available from the manufacturers shown will fill in the blank spaces. Some of these devices will be discussed in future columns.

These MMICs are not always the ultimate nor necessarily the desired choice for the front end of a low-noise receiver. However, they are good choices for second stages and input stages where noise figure is not at a premium. They are also recommended in intermediate frequency circuits or where gain is desired over a wide frequency range. Don't overlook their possiblities in transmitters as low-level stages. . . but that's another story!

We've only just scratched the surface. The MMIC manufacturers have already moved one major step forward. Not only are they providing amplifiers, but now whole "subsystems" are appearing — complete with rf amplifiers, mixer, local oscillator, and i-f amplifier, *all built on a single substrate!* Who knows what's in store in the future?

acknowledgments

I'd like to particularly thank Al Ward, WB5LUA, who sent me my first MMIC to "play" with. I have been hooked on them ever since.

---

new records

Summer is historically the time when VHF/UHF/SHFers go mountain-topping and otherwise take advantage of the improved propagation brought on by more favorable weather. This summer was no exception.

At least three difficult DX records have been recently set in North America. I don't yet have all the details, but here's a preliminary summary.

The 9-cm, 3456-MHz record now belongs to WA5TNY/5 and W7CNK/5 at approximately 222 miles. This was a tropo shot from Oklahoma City, Oklahoma, to Grapevine, Texas. Up in Oregon, WA3RMX and WB7UNU set a new 12-mm, 24.2-GHz line-of-sight record of approximately 115 miles. One

of the important facts about this record is that it was done with only 20 milliwatts and on two-way SSB!

Reports are circulating that some of the trans-Pacific VHF/UHF DX records may have also fallen in early August, especially on the 23-cm (1296-MHz) band between Hawaii and San Diego, California. I hope to be able to provide more details on these important events next month.

Congratulations to the new record holders!

references

1. Joe Reisert, W1JR, "VHF/UHF World: VHF/UHF Receivers," *ham radio*, March, 1984, page 46.
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6. Leonard D. Seader and James E. Starrett, "Unit Amplifier with Matched Input and Output Impedances," U.S. Patent No. 3,493,882, granted February 3, 1970.
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10. Robert G. Meyer et. al., "A Wide-Band Ultralinear Amplifier from 3 to 300 MHz," *IEEE Journal of Solid-State Circuits*, August, 1974, page 167.
11. Craig Snapp et. al., "Practical Silicon MMICs Challenge Hybrids," *Microwaves and rf*, November, 1982, page 83.
12. Jerry Hinshaw, N6JH, "Monolithic rf Amplifiers," *ham radio*, March, 1986, page 22.

important VHF/UHF events:

November 2	*Predicted peak of the Taurids Meteor Shower at 1520 UTC.*
November 3	*Predicted peak of the Casseopids Meteor Shower at 1515 UTC.*
November 4	*EME perigee*
November 17	*Predicted peak of the Leonids Meteor Shower at 0850 UTC.*
November 22/23	*ARRL EME Contest, second weekend.*
December 2	*EME perigee*
December 13	*Predicted peak of the Geminids Meteor Shower at 1250 UTC.*
December 21	*± 1 month. Winter peak of sporadic E propagation.*
December 22	*Predicted peak of the Ursids Meteor Shower at 0400 UTC.*
December 30	*EME perigee.*

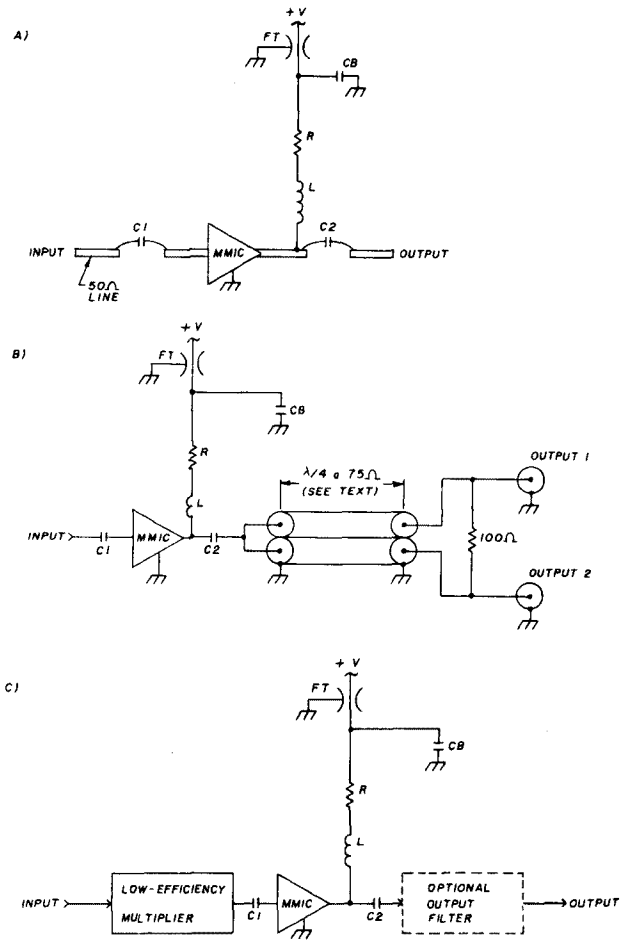


fig. 4. Examples of multipliers using MMICs. (A) is a frequency doubler. U1 and U2 are Avantek MSA0104s or equivalent. T1 is a trifilar wound transformer with three turns of No. 32 AWG per winding on a FairRite type 2843002402 or equivalent binocular core. Frequency range is limited by transformer characteristics. As shown the limit is about 300 MHz on the input. (B) is an example of a proposed tripler using MMICs. T1 and T2 are similar to T1 in *fig. 4A*. See text for further information on both circuits.

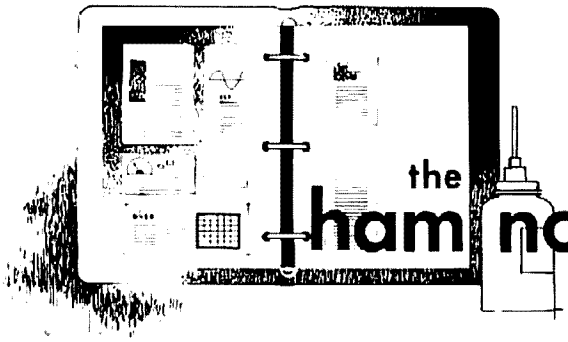
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waterproofing fittings

Keeping water out of antenna fittings, traps, and other outdoor assemblies can be a tough problem for most hams. Products often recommended in construction articles may be difficult to obtain. They may be expensive. But there is an alternative: Plastic Rubber™*, an inexpensive material that's readily available at most hardware counters and works better than a lot of other products I've heard recommended.

Plastic Rubber can be used freely on such things as coax fittings that will be exposed to the weather and on the seams of mini-boxes or other enclosures that must be kept watertight outdoors.

As it comes from the tube, it's either a gooey white or black liquid that quickly becomes tacky when exposed to air. It dries to the touch in 5 or 10 minutes, but remains soft and "tender" for an hour or two. When it has cured for several hours it dries to a tough, rubbery film a few mils thick. The film never becomes brittle.

I've used Plastic Rubber to seal coax fittings on 2-meter beams that remained in the air for several years of use, then lay on the roof for several more years of disuse after they were replaced by another beam. Even after all this exposure to Florida's summer sun, winter frosts, and plenty of thunderstorms, I found the seals intact. Two pairs of pliers had to be used to unscrew the shells of the PL-259s; the interior threads were as clean and uncorroded as they were when the fittings came from the store. In fact, it was difficult to peel the material from the metal. In many places, I had to scrape off the Plastic Rubber with a knife or file to get to the metal.

I haven't run tests on VHF or UHF losses in the material in those places where it's used as a dielectric, but any losses appear to be within acceptable limits. I can't detect any change in performance after application of the material to the solder connections of some SO-239s used to mount a half-wave balun on the beam.

Plastic Rubber also makes excellent seam-sealer for those leaky tents on a rainy Field Day. Apply it to the tent seams while the material is completely dry and let it cure for a day or two before re-folding the tent to prevent treated seams from adhering to each other. The material never stiffens and repeated folding of the tent doesn't erode the seal.

George L. Thurston, III, W4MLE

*"Plastic Rubber" is a trademark of the Duro Division of the Loctite Corporation of Cleveland, Ohio 44128.

ham radio

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3333LY	33 el	loop Yagi	902 MHz	18.5 dBi	$93

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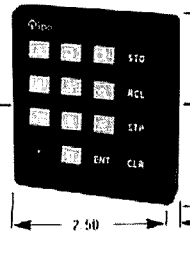
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high-performance receiver

Kenwood's new R-5000 high-performance receiver isn't just another all-band, all-mode receiver. The R-5000 covers 100 kHz-30 MHz in 30 bands, with additional coverage from 108-174 MHz available with installation of the optional VC-20 converter. It offers 100 memory channels, optional computer control, and much more.

Though similar in appearance to the TS-440S, the R-5000 matches the superior performance of the TS-440 receiver section. Its exclusive DynaMix™ direct mixing system ensures an honest 102-dB dynamic range (14 MHz, 500 Hz bandwidth, 50 kHz spacing.)

The R-5000 does not replace the popular high-performance R-2000, which is still available, but instead gives hams two high-performance receivers from which to choose.

For information, contact Trio-Kenwood Communications, P.O. Box 7065, Compton, California 90220.

ACC expands

Advanced Computer Controls, Inc., has moved its offices and manufacturing plant to a new facility in Santa Clara, California. The new building adds space for manufacturing, marketing, and engineering. (Northridge Square in Cupertino remains an acceptable mailing address.) ACC manufactures microcomputer-based control systems for Amateur Radio, commercial, and government radio users.

For more information, contact Advanced Computer Controls, Inc., 2356 Walsh Avenue, Santa Clara, California 95051.

Circle #309 on Reader Service Card.

new antenna from Mirage

The 1.2-44 LBX from Mirage utilizes the latest in ultra-high gain, low sidelobe design. The parasitic elements are pseudo-log tapered in length and spacing to produce wide usable bandwidth with low VSWR.

Because of its light weight, low windload and rugged construction, it's appropriate for use in higher gain multi-antenna arrays. Used alone it still delivers outstanding performance with better than 18.2 dBd gain across the specified bandwidth. This antenna requires no tuning; it's factory assembled and tested to insure that each unit meets or exceeds the advertised specifications.

For information, contact Mirage/KLM Communications Equipment Inc., P.O. Box 1000, Morgan Hill, California 95037.

Circle #301 on Reader Service Card.

two new communications microphones

Shure Brothers Inc., Evanston, Illinois, has added two new, economical models to its line of communications microphones.

The Shure 550L Base Station Microphone, suitable for radio communications, paging, and dispatching system applications, shares many design features with Shure's well-known Model 450k. Its omnidirectional cartridge is specially tailored for voice intelligibility, and its balanced, low-impedance design makes it useful for long cable runs and use under severe hum conditions.

Constructed of strong, corrosion-proof ARMO-DUR™, it features Shure's long-life "Million Cycle" press-to-talk bar switch. User Net Price is $66.25.

The Shure Prologue 6L Handheld Communications Microphone — the lowest-priced handheld communications microphone ever offered by Shure — features a dynamic low-impedance cartridge with tailored response for high intelligibility, a sturdy push-to-talk switch, a durable coiled cable, an extra-strength mounting bracket, a relay closure circuit, and compact, lightweight construction. The Shure Prologue 6L's list price is $41.75.

For further information about the Shure 550L and Prologue 6L, contact Shure Brothers Inc., Customer Services Department, 222 Hartrey Avenue, Evanston, Illinois 6020-3696.

Circle #308 on Reader Service Card.

test monitor

Comtest Systems has introduced the Model 3000B Communications Service Monitor. Said to be the only monitor in its size and price range that gives the kind of testing capability usually available only in larger instruments costing thousands more, this lightweight, versatile instrument provides three different displays — waveform CRT, meter, and digital counter.

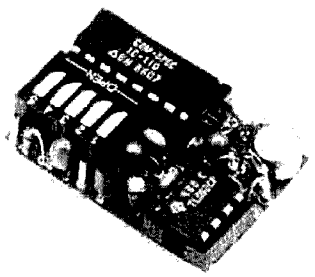
Receive features include a tuneable range of 400 kHz to 1000 MHz, combining high sensitivity with good intermod rejection and wide frequency range for "off the air" monitoring. A high-resolution audio frequency counter LCD display assures versatility and accuracy. The counter also permits the display of CTCSS frequencies directly off the air. The large CRT provides instantaneous information about modulation waveforms and internally generated markers can be displayed at ± 5 kHz or ± 600 Hz for "at a glance" identification of FM modulation peaks.

For further information, contact Comtest Systems, Inc., P.O. Box 470, Beech Grove, Indiana 46107.

Circle #307 on Reader Service Card.

programmable encoder

Communications Specialists has announced the introduction of its new programmable encoder, Model SS-32P. The SS-32P allows the user to specify the available tones in the memory. Up to 32 tone frequencies are stored in an EPROM memory; these tones may be standard or non-standard, and may be changed. The working tone is retrieved from the memory with a five-position DIP switch mounted on the circuit board.



The SS-32P is available in either CTCSS or burst tone format: Model SS-32PA will operate on any tone from 67.0 to 250.0 Hz; Model SS-32PB will operate between 250.0 and 3000 Hz. Both versions measure 0.9 x 1.3 x 0.4 inches and will operate on voltages as low as 6 VDC for handheld applications and up to 25 VDC for mobile or base station use.

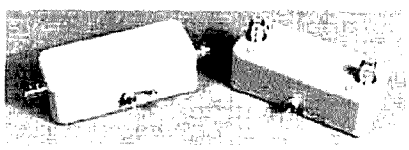
The SS-32P allows the use of non-standard tones for extra flexibility on crowded channels and for added security on tone controlled repeaters. It also simplifies multiple tone switching (up to six tones!).

Priced at $28.95, the SS-32P is covered by a one-year warranty and is available for immediate delivery from stock. For information and a free catalog, contact Communictions Specialists, Inc., 426 West Taft Avenue, Orange, California 92665-4296.

Circle #312 on Reader Service Card.

audio/video amplifier

An audio/video amplifier from WI-COMM Electronics, Inc., operates over the frequency range from 300 Hz to over 10 MHz and features multistage design, with gain rated at 32 dB; noise figure, 3 dB; 3rd order intercept point, 33 dBM; power output, 0.10 watt into 200 ohms; input/



output VSWR 2:1; and reverse isolation, 55 dB. Standard connectors are BNC female, 26 VDC. Reverse polarity protection is included. Possible applications include audio/stereo, SCA, AM stereo, video, digital audio, and LF through HF communication links.

For details, contact WI COMM Electronics, Inc., P.O. Box 5174, Massena, New York 13662.

Circle #311 on Reader Service Card.

NEW products

new Heath catalog

Heath Company of Benton Harbor, Michigan, has announced the release of its upated fall catalog. With an expanded line of starter kits, computers, weather instruments, and a variety of electronic products, the redesigned Heathkit catalog brings a new, refreshing look to its pages.

Heathkit's expanded computer software/hardware line includes the new HyperACCESS communication software, designed to permit communication with other microcomputers and databases. Also included is the new Teach-N-Tutor Authoring software from Zenith Data Systems, which allows programmers to create tutorials for computer-aided instruction (CAI). New computer hardware includes Epson printers and a 2400-Baud modem from US Robotics.

For a free copy of the catalog, contact Heath Company, Dept. 150-785, Benton Harbor, Michigan 49085. In Canada, contact Heath Company 1020 Islington Avenue, Dept. 3100, Toronto, Ontario M8Z 5Z3.

Circle #310 on Reader Service Card.

New Larsen Antennas

The new YA5 series of Yagi directional antennas includes the YA5-900, designed to operate in the 900-960 MHz range. It features all-weather constuction (i.e., coax termination *inside* the Yagi – away from wind and rain – plus a black weather-resistant coating. Other features include user interface via N-male connections to the N-female cable-mounted connector supplied with the antenna kit. The mounting hardware (double-welded, heavy-duty U-bolts) will fit any pipe from 1-1/8 to 2-1/2 inches in diameter.

Specifications include 10 dBd gain, 45-degree horizontal beamwidth, 55-degree (estimated) vertical beamwidth, and maximum power of 300 watts.

For more information, contact Larsen Electronics, P.O. Box 1799, Vancouver, Washington 98668.

Circle #313 on Reader Service Card.

eight-pole crystal filters

Several new eight-pole crystal filters are available from International Radio.

The new Kenwood TS-440 8-pole crystal filter package consists of SSB 2.1-kHz or 1.8-kHz and 400 Hz or 250 Hz filters. An SSB 2.1 kHz matched crystal filter set is also available. The typical response of the matched set is 2.1 kHz at 6 dB and 2.8 kHz or less at 60 dB.

New filters are also available for ICOM units. Stock No. IRI-455H400X is a CW, 400-Hz (eight-pole) that serves as an exact replacement for the FL-52A. It fits the IC-751(A), the IC745, the IC740, etc.

Stock No. IRI-455H1.2X is an eight-pole SSB, 2.4 kHz exact replacement for the FL-44A. It fits the IC730, IC740, IC745, and the R70 and R71.

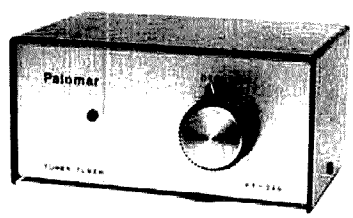
The typical shape factor for the SSB filters is 1.66 or less; for CW filters, the center frequency is 8.0 or 9.0 MHz = 3.2 or less. The 455 kHz center frequency is 2.0 or less. All International Radio crystal filters are guaranteed to the original purchaser for two years.

Many other ICOM crystal filters available. For details, contact International Radio, Inc., 747 South Macedo Blvd., Port St. Lucie, Florida 33452 (enclose SASE).

Circle #314 on Reader Service Card.

Palomar "Tuner-Tuner"™

Palomar Engineers has announced a new operating aid, the Tuner-Tuner.™ It connects between your transceiver and antenna tuner, and allows you to tune your tuner precisely *without transmitting*. The built-in 50-ohm noise bridge gives an audible null in the receiver when the tuner matches the coax line to 1:1 SWR. Your transmitter never need operate into high SWR, and you need not cause interference on the band while tuning up.

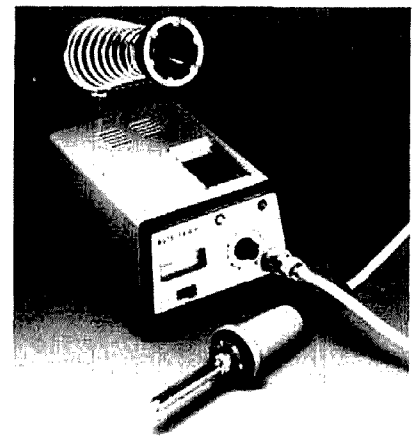


The Model PT-340 Tuner-Tuner is priced at $99.95 plus $4 shipping. For more information, contact Palomar Engineers, Box 455, Escondido California 92025.

Circle #315 on Reader Service Card.

temperature-controlled soldering station

Start Manufacturing has announced the release of a versatile soldering station that permits the interchanging of 30, 48, and 60-watt irons, thus providing the operator with three stations in one. The station also features a unique control that keeps the temperature to within ±8 degrees F of the setting. It also offers direct temperature readout, easy and fast calibration, and a feature that permits the station to be set and locked at a specific temperature. The station comes with a standard 48-watt pencil iron and adjusts over the full temperature range of 200 through 900 degrees F without having to change the tip or heating element.



For information, contact Start Manufacturing, Inc., 15775 North Hillcrest, Suite 508, Dallas, Texas 75248.

Circle #306 on Reader Service Card.

two-band rotary dipole

The CD 78 from Orion Hi-Tech is a 80- and 75- meter rotary dipole. Approximately 40 percent shorter than a full-sized dipole, its unique design results in a less than 2 dB reduction in radiation efficiency over a full-size antenna.

The CD78 is equipped with a special high-performance matching unit that is used to switch the antenna between the two bands. 2:1 VSWR bandwidth is approximately 50 kHz on both 80 and 75.

Built from high-strength, heavy-duty aluminum, it features a very strong element-to-boom mount. All hardware is hot-dipped galvanized after fabrication. The CD78 will handle full legal power and is rated at 40 m/s wind survival.

For details, contact Orion Hi-Tech, P.O. Box 8771, Calabasas, California 91302.

Circle #316 on Reader Service Card.

free QSL cards!

Varta Batteries, Inc. – one of the world's three largest battery manufacturers and a leading U.S. supplier of batteries for short-wave radio, photography, data processing, and other high-tech applications – is offering a package of 20 QSL cards to Amateur Radio Operators, free upon request.

The postcards are printed in full color with an illustration of world map as a background for Varta's wide line of batteries for shortwave Amateurs. The address side of the card follows the typical QSL card form.

Readers who would like to obtain a package of cards should send their name and address to: Paul Silliman, Varta Batteries Inc., 300 Executive Blvd. Elmsford, New York 10523-1202.

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R-2000 repeater controller

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NEW products

R-2000 repeater controller, which can control a complete repeater system including two remote bases and a control receiver. The fully-programmable, microprocessor-based controller includes many features for today's demanding repeater service. Features include autopatch with memory dialer, ALC and call progress detection, mailbox/bulletin board, and dual RS-232 ports for interfacing to a computer or printer.

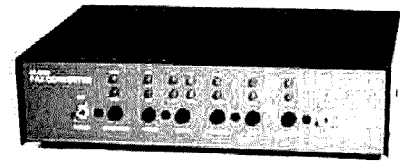
For more information, contact Resultant Engineering, 175 East Homestead, Sunnyvale, California 94087.

Circle #317 on Reader Service Card.

FAX unit prints photos

Universal Shortwave has introduced the microprocessor-controlled DES M-800 FAX Converter, which allows reception and printing of pictures, photos, maps, and marine and National Weather Service information, as well as satellite earth imagery and world meteorological charts transmitted by shortwave and satellites.

The M-800 Facsimile Converter connects to the audio output of any quality communications receiver or satellite receiver (AM or FM). Audio



is converted to graphics data and printed out on most Epson parallel dot-matrix computer printers (FX-85, LQ-800, or equivalent). Print size is 8 inches by length as needed. Aspect ratio is 1:1.

FAX photos are printed on plain paper in 16 levels of gray scale. Weather maps (NWS), charts, and other black-and-white transmissions yield exceptional clarity.

For details, contact Universal Shortwave, 1280 Aida Drive, Reynoldsburg, Ohio 43068.

Circle #320 on Reader Service Card.

D2M2

Known as D2M2™ —*not* DMM — because they offer twice the range and resolution and twice the functions of ordinary DMMs with 2000-count capability, the new 4000-count 4-digit multimeters with 40-segment analog bar graph from North American SOAR are priced to rival ordinary 2000-count DMMs.

All D2M2 range crossover levels are at the 4, 40, 400, and 4000 points to allow measurements of 3.999 or 39.99, etc. Accuracy is better than 0.3 percent for Models 4020, 4030, and 4040, and better than 0.1 percent for Model 4050. All models have been constructed to be shockproof and dustproof for operation in harsh environments.

All models are auto- and manual-ranging with Data Hold Function, Continuity Beeper, Diode Test, and more. Selectable models feature illuminated LCD, Frequency Counter, Relative Set, Memory, Temperature and Min/Max Indication.

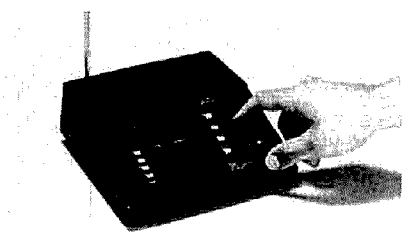
Optional adapters expand functions to include Capacitor Test, High AC/DC current, Transistor GAIN and Loss Test, and Temperature. Full overload potection includes two fuses plus transient noise to 6000 volts. 1500-hour battery life is typical with two "AAA" alkaline cells and auto power down. All D2M2 units are supplied with batteries, safety leads, spare fuses and 3 year limited warranty, calibration accuracy guaranteed for 1 year.

Prices for D2M2 Series meters in single-lot quantities are as follows: Model 4020, $99.99; Model 4030, $139.00; Model 4040, $159.00; and Model 4050, $179.00, with delivery from stock. For full specifications or additional information, contact North American Soar Corporation, 1126 Cornell Avenue, Cherry Hill, New Jersey 08002.

Circle #319 on Reader Service Card.

15-channel programmable scanner

Automatic search, scan delay and a priority channel are features that are normally reserved for expensive, top-of-the-line scanners. But one modestly priced programmable scanner, the Regency R1075, is packed with all of these features. Fully programmable, the 15-channel scanner can receive more than 15,000 frequencies from six of the most popular public service bands.



Designed for the beginning scanning enthusiast, as well as the veteran seeking a reliable back-up unit, the scanner can be programmed to search a frequency range for active new frequencies. With its priority channel and scan delay functions, the scanner keeps listeners from missing important transmissions. When activated, the priority channel automatically overrides all other calls so that broadcasts from a favorite channel are never missed. Scan delay puts a 2-second pause at the end of a transmission so that calls and answers can be heard before the scanner resumes its scanning cycle.

Priced at $179.95, the scanner covers six full bands, including VHF-Low (30-50MHz), VHF-Amateur (144-148 MHz), VHF-High (148-174 MHz), UHF Amateur (440-450 MHz), UHF (450-470 MHz), and UHF-T (470-512 MHz).

For information, contact Regency Electronics Inc., 7707 Records Street, Indianapolis, Indiana 46226.

Circle #302 on Reader Service Card.

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IC-02AT, 04AT
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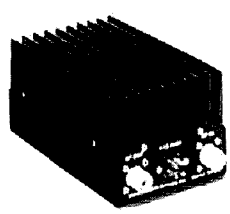
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NEW products

RFI technical note

Palomar Engineers has published a helpful note titled ""Using Ferrite Beads to Keep rf Out ol TV Sites, VCR'S, Burglar Alarms and Other Electronic Equipment." The note explains what kinds of beads and toroids to use and where to put them to suppress most common rf interference problems.

For a free copy, contact Palomar Engineers, Box 455, Escondido, California 92025.

Circle #303 on Reader Service Card.

oscilloscope frequency extenders

Radio Engineers has recently introduced several new products including two models of frequency extenders for the home builder with an oscilloscope that can ""see" only up to 5 MHz. Designated the Type HFX-1 and HFX-2 Oscilloscope Frequency Extenders, these units allow 5 MHz or 10 MHz bandwidth oscilloscopes to see rf signals as high as 55 MHz. Type HFX-1 contains an rf signal source and a broadband mixing circuit that converts the high frequency signal to an intermediate frequency within the oscilloscope's range. Type HFX-2 is designed to be used with an external signal generator but still gives the same frequency coverage. Both units have 50-ohm input impedances and utilize BNC connectors to interface to the circuit under test and the oscilloscope. Also available is Type DBX-30, a fixed 30-dB attenuator that allows the extenders to sample higher power levels. A high-impedance transformer to be used at the input of the oscilloscope extenders is planned for release in the near future.

Data sheets are available on these and other products. For copies, contact Radio Engineers, 3941 Mount Brundage Avenue, San Diego, California 92111.

Circle #304 on Reader Service Card.

turnable preamplifier antenna

Ameco's Model TPA is a dual-function unit that can be used as a preamplifier to improve the gain of a receiver or as an indoor active antenna when an outdoor antenna isn't available. It contains a tuned rf amplifier that covers all frequencies from 0.22 to 30 MHz, including Amateur bands, all foreign broadcast bands, citizen's band and all other services within this range. A dual-gate FET provides an excellent noise figure and over 20 dB gain. The weak-signal performance of most receivers is improved.

Priced at $74.95, Model TPA uses either an internal 9-volt battery or an ac adapter such as Ameco's Model P-9T. As a preamplifier, the input matches most antennas. Long wire, 300 ohm and random-length antennas can also be used with good results.

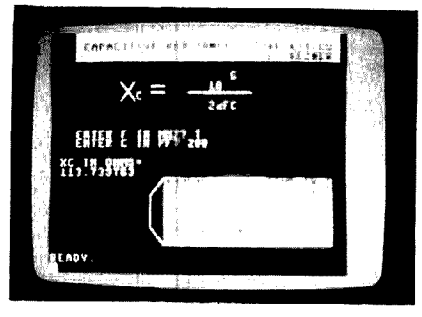


For information, contact Ameco Equipment Company, a Division of Ameco Publishing Corporation, 220 East Jericho Turnpike, Mineola, New York 11501.

Circle #322 on Reader Service Card.

1.0 Formula Disk

RF Kit Company has announced a new software disk for the Commodore C-64. This new release includes many formulas such as Ohm's law, inductive and capacitive reactance, resonant frequency, dipole antennas, temperature conversions from C to F and F to C, Great Circle bearings and distances to anywhere, Tower Stress calculations, Noon Meridian passage lor



True North antenna adjustments, how to wind coils lor a specific inductance, how to make a meter shunt and how to program your own favorite formulas on the disk. The RF Kit 1.0 disk sells for $14.95 plus $1.00 postage and packing.

For information, contact RF Kit Company, P.O. Box 27127, Seattle, Washington 98125.

Circle #321 on Reader Service Card.

Communications Satellites

Now in its expanded second edition, *Communications Satellites* covers spy and surveillance satellites, U.S. and Russian manned space missions, military, tactical and scientific satellites, oceanographic and weather orbiters, deep space probes, navigational and communications satellites, and private and direct broadcast satellites. This directory of space communications includes chapters on channelization band plans, transponder identification, international satellites, and an informative history of earth satellite development.

Priced at $14.95 (plus $3.50 for shipping and handling), the book may be ordered from Ham Radio's Bookstore, Greenville, New Hampshire 03048.

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DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

PACKET PROGRAM FOR YOUR PC. A telecommunications program especially written to interface your packet controller with your PC or compatible. Features Split Screens, ASCII, Binary, X Modem file transfers, DOS shell, Macro Keys, 5 types time date oper ID. Lots of goodies. Software support. Write for complete information. VISA/MC. Kalt and Associates, 2440 E. Tudor Rd, Suite #138, Anchorage, AK 99507. (907) 248-0133.

BACK ISSUES HR Oct 71 thru Dec 81, 72 77 in Binders, $60 US. VE3BJY, B. Brenko, 2 Vista Drive, Fonthill, Ont. L0S 1E0.

WANTED: National NCX-500 transceiver and matching AC-500 power supply or National NCX-1000 transceiver. 512 647-0748, K3WB, 5302 Charter Oak, San Antonio, TX 78229.

4-1000A PULLS @70.00 Raytheon 4D32 @35.00 ea. 4CX50B @20.00. Collins plate transformer for KWS1 75.00, heavy duty TV power transformers new boxed 20.00 ea, HT-37, @100.00. Eldico 100F exciter 150.00. White porcelain beehive insulators @1.00. Hard to find receiving tubes, low prices. Inquire. M. Levy, W5OJT, 101 East Driftwood #44, FBG, TX 78624. Tel (512) 997-2534.

SPEECH SYNTHESIZER KIT FOR C-64 uses allophones to obtain virtually unlimited vocabulary. P.C. board, all parts, instructions $30.00 plus $2.00 UPS. Dave Roscoe, W1DWZ, 49 Cedar Street, East Bridgewater, MA 02333.

ORIGINAL ISSUES OF HAM RADIO March 1986 (first issue) through June 1974. Complete. Good condition. Would prefer to sell as a complete set but will sell individual issues if offer is right. Make offer. KX9P, 2675 Clark Avenue, Marion, Iowa 52302.

$$$$$SUPER SAVINGS on electronic parts, components, supplies and computer accessories. Free 40 page catalog for SASE. Get on our mailing list. BCD ELECTRO, PO Box 830119, Richardson, TX 75083 or call (214) 690-1102.

HOMEBREW PROJECTS. List SASE. WB2EUF, PO Box 708, East Hampton, NY 11937.

DYNAMIC COLOR NEWS: Monthly newsletter provides programming instruction for Radio Shack Color Computers. Includes HAM RADIO section. FREE sample. Dynamic Electronics, Box 896, Hartselle, AL 35640. (205) 773-2758.

SLEP MAIL ORDER SPECIALS HP/Boonton 202H AM/FM signal generator, 54 MHz thru 216 MHz $275.00. Military small portable URM-25D signal generator 10 kHz thru 50 MHz $245.00. Military TS-403/U microwave signal generator 1.8 GHz thru 4.2 GHz $285.00. Military OS-8C/U small portable 3" oscilloscope, ideal for bench work or modulation/RTTY testing $59.50. HP606A signal generator 50 kHz thru 65 MHz $375.00. HP608E signal generator 10 MHz thru 480 MHz $445.00. HP614 microwave signal generator 900 MHz thru 2100 MHz $345.00. All lab calibrated. Satisfaction guaranteed. Have quantity, VISA, M/C or check. Add shipping. Phone Bill Slep (704) 524-7519. Slep Electronics, Highway 441, Otto, NC 28763.

CABLE TV CONVERTERS & EQUIPMENT: Plans and parts. Build or buy. SASE for information. C & D Electronics, PO Box 1402, Dept. HR, Hope, AR 71801.

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R-390A Receiver: $195 checked; $115 reparable. Parts, tubes, sections. Info SASE. CPRC-26 six meter transceiver (see HR, March 1985) $17.50 apiece, $32.50 pair (add $4.50 unit shipping). Baytronics, Box 591, Sandusky, OH 44870. 419-627-0460 evenings.

MARCO: Medical Amateur Radio Council, Ltd. operates daily and Sunday nets. Medically oriented Amateurs (physicians, dentists, veterinarians, nurses, physiotherapists, lab technicians, etc.) invited to join. Presently over 550 members. For information write MARCO, Box 73's, Acme, PA 15610.

IBM-PC RTTY/CW. New CompRtty II is the complete RTTY/CW program for IBM-PC's and compatibles. Now with larger buffers, better support for packet units, pictures, much more. Virtually any speed ASCII, BAUDOT, CW. Text entry via built in screen editor! Adjustable split screen display. Instant mode/speed change. Hardcopy, diskcopy, break-in buffer, select calling, text file transfer, customizable full screen logging, 24 programmable 1000 character messages. Ideal for MARS and traffic handling. Requires 256k PC or AT compatible, serial port, RS-232C TU. $65. Send call letters (including MARS) with order. David A. Rice, KC2HO, 25 Village View Bluff, Ballston Lake, NY 12019.

ELECTRONIC ENCLOSURES: Attractive, reasonably priced. Aluminum top and sides. Walnut stained solid wood ends. Variety of sizes and custom sizes available. Sample $5.00. Information free. Energy Engineering, Custom Division, Rt. 4, Box 330, Fayetteville, AR 72701.

NATIONAL RADIO EQUIPMENT manual list or NCL 2000 parts kits. SASE. Maximilian Fuchs, 11 Plymouth Lane, Swampscott, MA 01907.

ATTENTION AMATEURS Send for Free discount catalog. Amateur Communications, 2317 Vance Jackson, San Antonio, TX 78213. (513) 734-7793.

8877 VHF AMP KITS: HV power supplies, CX600N relays, Mu-Tek LTD front end boards for IC251, IC271, EME newsletter and QRO parts. SASE for new catalog. KB7Q, "Q" Products, 417 Staudaher Street, Bozeman, MT 59715.

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CHASSIS and cabinet kits. SASE K3IWK, 5120 Harmony Grove Road, Dover, PA 17315.

COMING EVENTS

Activities — "Places to go . . ."

CONNECTICUT: November 9. SCARA annual indoor Flea Market, North Haven Park and Recreation Center, North Haven. Sellers 7 AM. Buyers 9 AM to 3 PM. Tables $10.00 advance, $15.00 at the door. Buyers $2.00. For information or table reservations SASE with phone number to SCARA Flea Market, PO Box 81, North Haven, CT 06473. Reservations must be received by November 3, 1986. No reservations taken by phone. For information ONLY contact Brad (203) 265-6478, 7 PM to 10 PM.

MASSACHUSETTS: November 8. New England annual DXCC Banquet. Masonic Lodge Building, Monument Square, in historic Concord Center. Afternoon session 2 PM to 5:30 PM. Banquet 6:30 PM. For further information: Charles Lukas, Jr., W1DOH, RFD 1, 24 Durkee Road, Acton, MA 01720. (617) 263-3743.

MICHIGAN. November 30. The Oak Park High School Electronics Club presents the 17th annual Swap 'N Shop, Thanksgiving Sunday. Oak Park High School, Oak Park, MI. Donation $2.00. 8' tables $8.00.

MICHIGAN: November 2. The Oak Park ARC will hold its largest ever 1986 SWAP 'N SHOP. New location: City of Southfield Civic Pavilion, Evergreen Road between 10 and 11 Mile Roads, northwest Detroit suburb of Southfield. Ham Radio and Computer activities from 9 AM to 5 PM in the new fully carpeted 30,000 sq. ft. pavilion. The Detroit Area Repeater Team (DART) will provide food and refreshments. Admission $4, under 12 free. VE3's at par. Tables $10.00. Advanced reservations required. Talk in on DART 146.04/.64 and 146.52 simplex. For further information SASE to OPARC Swap 'N Shop, 303 South Vermont Avenue, Royal Oak, MI 48067. Swap 'N Shop Hotline (313) 399-3991.

CALIFORNIA: FCC exams, Novice-Extra. Sunnyvale VEC ARC. (408) 255-9000 24 hour. 73, Gordon, W6NLG. VEC.

MASSACHUSETTS: The MIT UHF Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday, November 19, 7 PM, MIT Room 1-134, 77 Mass Avenue, Cambridge MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-0160/646-1641 or Craige Rodgers (617) 494-1986. Exam fee $4.25. Bring copy of current license (if any), two forms of picture ID and completed form 610 available from FCC in Boston (223-6609).

TEXAS: November 7-9. AMSAT will hold its 4th annual Space Symposium and Annual Meeting, Dallas/Fort Worth Airport Hilton Hotel. Speakers include experts from around the world addressing the latest in OSCAR. Featured speaker Dr. Martin Davidoff, K2UBC. Additional details and registration information available from AMSAT HQ. (301) 589-6062.

MASSACHUSETTS: November 22. The Honeywell 1200 Radio Club, sponsor of 147.72/12 repeater and the Waltham Amateur Radio Association, sponsor of 146.04/64 repeater, will hold their annual Amateur Radio and Electronics auction, Honeywell Plant, 300 Concord Road, Billerica. Doors open 10 AM. Free admission and parking. Snack bar and bargain parts store. Talk in on both repeaters. For more information: Doug Purdy, N1BUB, 3 Visco Road, Burlington, MA 01803.

OHIO: November 16. The Massillon ARC will sponsor "Auctionfest 86", Massillon K of C Hall off Route 21. 8 AM to 5 PM. Sellers setup 7 AM. Admission $3.50 advance; $4.00 at the door. 8' tables $7.00. Free parking. Refreshments available. Auction starts 11 AM. Talk in on W8NP, 147.78/.18. For advance registration and information: MARC, PO Box 73, Massillon, Ohio 44646. Please SASE.

INDIANA: November 9. The allen County Amateur radio Technical Society will prsent the 14th annual Fort Wayne Hamfest. Allen County Memorial Coliseum, Coliseum Blvd, US 30. Doors open 8 AM to 4 PM. General admission $3.50 advance; $4.00 at the door. Children 11 and under free. Tables $10.00 each. AC power extra. Premium tables $25.00 each. Non-ham activities. Banquet Saturday night. Nearby hotels and motels. VE exams Saturday, November 8 by advance registration only. kTalk in on 146.28/.88. Forinformation or reservations: AC-ARTS Hamfest, PO Box 10342, Fort Wayne, IN 46851. For information ONLY Bernie Holm, K9JDF, Hamfest Chairman, (219) 485-0164, 6-10 PM EST.

WISCONSIN: November 15. The Milwaukee Repeater Club is sponsoring the 2nd annual 6.91 FRIENDLY FEST. Eagle's Club, 24th and Wisconsin Avenue. 8 AM TO 1 PM. Sellers 7 AM. Tickets $3.00. 4' tables $4.00. Save $1.00 by sending SASE with payment to Milwaukee Repeater Club, PO Box 2123, Milwaukee, WI 53201 before 11/8/86. Talk in on 146.91 and 146.52.

MINNESOTA: December 6. Annual Handi-Ham Winter Hamfest. The Eagles Club, Faribault. Registration 9 AM. Handi-Ham equipment auction. Dinner at noon followed by a program. For more information contact Don Franz, W0FIT, 1114 Frank Avenue, Albert Lea, MN 56007.

OPERATING EVENTS

"Things to do . . ."

November 9-11: A.F.A.R., the Armored Forces Amateur Radio Net will operate a special event station, 1700 UTC 11/9 to 2400 UTC 11/11. All bands, #10 SASE to WB1DWR, 16 Berkeley Circle, Newington, CT 06111.

December 6: The University of Idaho ARC, W7UQ, will hold its 2nd annual Alumni Reunion on the Air, 2000Z, December 6 to 0400Z December 7. All Amateurs, especially U of I alumni, are invited to participate. Listen for "CO Reunion". QSL available by sending SASE via callbook address. For more information contact W7UQ.

November 9. In observance of Veteran's week, members of the Hamfesters Radio Club, Chicago, will operate from the Hines VA Hospital's Robert K. "Pappy" Wade, K9CDH Memorial Ham Shack using Hine's club call K9WFN. 1500Z to 0300Z. 40, 20, 2 meters FM and USB. Send QSL, QSO number and 9x12 SASE with 39 cents postage to Hamfesters Radio Club, Inc., Chicago, c/o Robert K. "Pappy" Wade Memorial Ham Shack, Bld 8, Hines Veterans Administration Hospital, Hines, Illinois 60141.

November 15. The Laurel, MD ARC will operate K3LDE from 1300Z to 2300Z for the runing of the Washington International Horse Race. Send QSL, contact number and #10 SASE for certificate to: LARC, 1120 12th Street, Laurel, MD 20707.

W.E.C.A. (Westchester Emergency Communications Association) initiates Equipment/"Elmer" Banks to help new hams get on the air. More information write the club at PO Box 131, North Tarrytown, NY 10591.

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AMATEUR RADIO RESOURCES DIRECTORY 1986-87

Have a question that no one can answer????? "Fred" will. The "white" pages list the folks who can answer any Amateur Radio related problems or question you might have: ARRL Directors, Vice Directors, Assistants, Advisory Committee members, Field volunteers, VEC Volunteer examiners, all organized geographically by ARRL Division. The "Blue" pages contain a QST 10 year cumulative index, QEX and Gateway bibliographies, TIS info and more! Every ham should have a copy of this book in their shack. 1986

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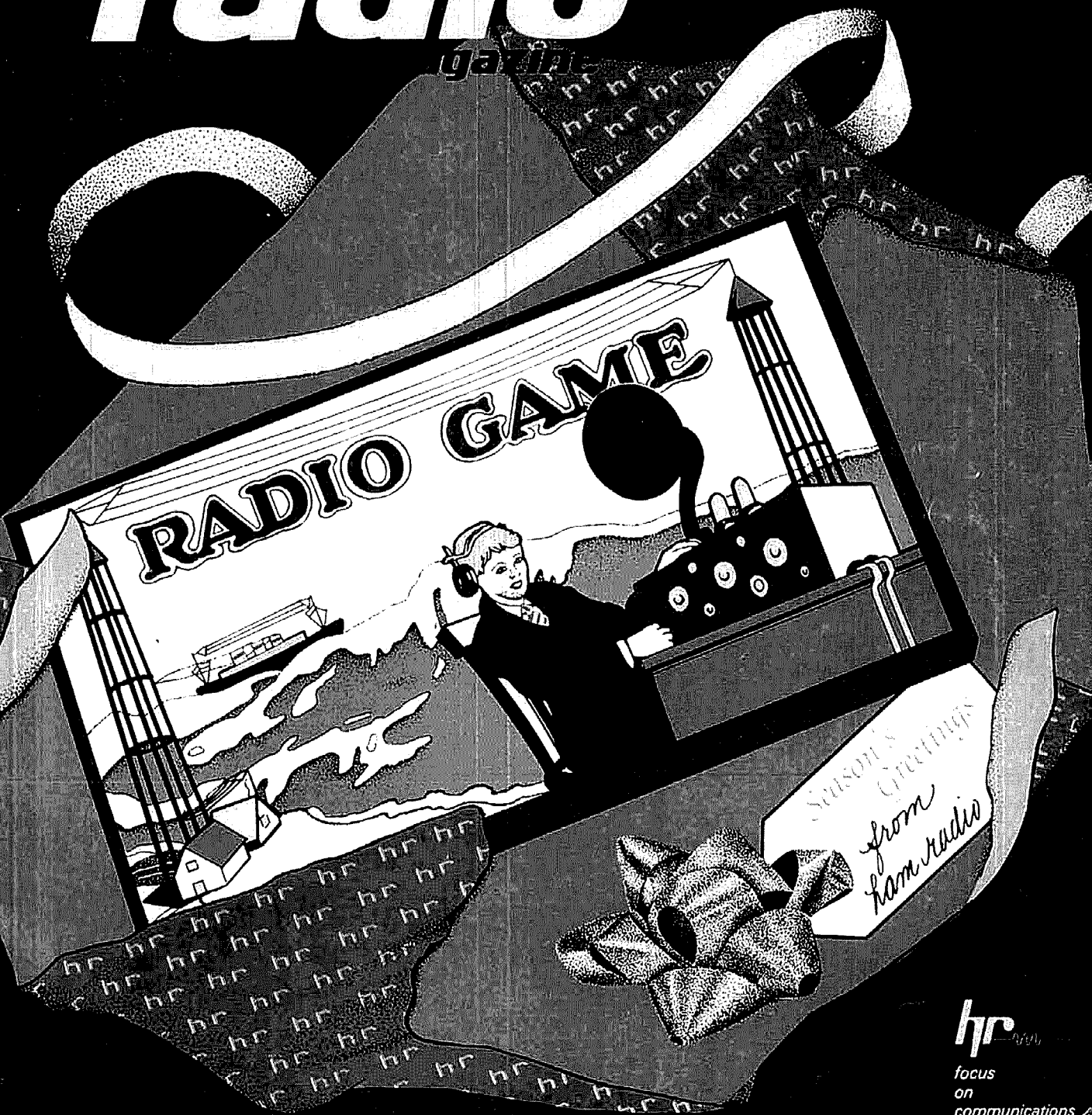
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publisher

Rich Rosen, K2RR
editor-in-chief
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Dorothy Rosa, KA1LBO/KT
assistant editor

Joseph J. Schroeder, W9JUV
Alfred Wilson, W6NIF
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director of advertising sales

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advertising production manager

Susan Shorrock
circulation manager

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circulation

cover art:
Barbara Smullen

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REFLECTIONS

13 exciting years: looking back

It's been exactly 13 years since Skip Tenney suggested that Amateurs might welcome a short-lead time "newsletter" to cover current news pertinent to Amateur Radio. Skip, Jim Fisk, and I then set to work, and just over a month later the first issue of "HR Report," dated January 2, 1974, was handed out at the SAROC Convention in Las Vegas.

Needless to say, Skip was right. Thirteen years later, *HR Report* is still going strong as *Westlink Report*, under the direction of Bill Pasternak and Burt Hicks. And today, despite a couple of *HR Report's* early would-be competitors having dropped by the wayside, the Amateur fraternity supports two additional broad-coverage newsletters plus a host of others devoted to one special interest or another.

Reviewing the past 13 years through the pages of *HR Report*, *Westlink Report*, and *Presstop* (*ham radio's* newspage, originally distilled from *HR Report but independently written and produced since 1982), one is tempted to say that these have to have indeed been the most exciting 13 years in the history of Amateur Radio. Just consider these Amateur Radio happenings reported in HR Report* during its first two years:

The first thing you'll notice is that while some things change a lot, others change very little or not at all. In the very first *Ham Radio Report*, we reported that CB was threatening 220 MHz and the Emergency Medical Service was after part of 70 cm, FCC license processing was taking 8 to 10 weeks, and a planned SY5MA Mt. Athos operation had been scrubbed. Later that first year *Ham Radio Report* also reported on the initial efforts that led to our getting new bands at the 1979 World Administrative Radio Conference, licensing of the first ATV repeater, FCC elimination of log-keeping requirements, the launch of OSCAR 7 in November, and in December, the FCC's proposed "top-to-bottom" restructuring of the Amateur Service (Docket 20282).

1975 saw Amateur and CB license fees reduced, "environmental impact" and OSHA became Amateur concerns, and the FCC ban on 27-MHz (but not "all-band") linears (and the quick introduction of 4-watt drive all-band "Amateur" linears by various CB suppliers). Dick Baldwin, W1RU, replaced John Huntoon, W1RW, as ARRL General Manager, and the FCC proposed an Automatic Transmitter Identification System for all 25-960 MHz transmitters (except Amateur). The WARC-79/WG/AM (Working Group/Amateur Radio) had its first meeting in May, the FCC reported CB license applications arriving at 15,000 per working day, off-shore navigation systems encroached on 70 cm, the FCC OK'ed special events callsigns, attendance at Dayton exceeded 10,000, *QST* and *ham radio* (and later *73*) announced a change to 8 1/2 x 11-inch formats effective January, 1976, and Texas proposed restricting all rf radiation. John Johnston (then K3BNS, now W3BE) became Chief of the FCC's Amateur and Citizens Division and announced that his would be a "deregulatory" policy (which began by greatly reducing repeater license paperwork), new specific mm-wave Amateur bands through 250 GHz were established, and Amateurs got a nice Christmas present from the FCC when it dropped the minimum operating time and code proficiency requirements for license renewal.

Each succeeding year records an equally diverse list of happenings. Here's just a sampling of some of Amateur Radio's exciting events that occurred during that time:

In space, AMSAT's relatively primitive (but effective) OSCARs 6, 7, and then 8 were joined by the comparable Russian RS birds, and later on, by the very sophisticated OSCAR 10 (whose elliptical orbit finally made worldwide Amateur satellite communication possible) and Japan's "Fuji." Amateurs themselves finally got on the air from space when W5LFL and W0ORE operated from the Space Shuttle.

The U.S. Amateur Service itself grew at a fantastic rate, at least during the CB boom years. At the end of 1974 there were only 253,357 licensed Amateur operators in the United States. Today there are more than 421,000! Much of that growth paralleled the CB boom, which at its peak had users variously estimated at from 15 to 30 million. From 1975 through 1982, an average of over 20,000 new Amateurs were licensed each year, but that died quickly to the present thousand or so a year after CB lost its appeal. The FCC's "restructuring" proposal, which included a no-code "Communicator" class license, also died when the League found strong opposition among its membership.

At Geneva we did get the three new hf bands our volunteer group had worked so hard for, despite the warnings of "gloom-and-doom" advocates who claimed we'd be lucky to come out of WARC '79 with any Amateur bands at all. All the Amateur monthlies increased not only size but price — when HRR began, all had cover prices under a dollar; today, all cost between $2.50 and $3.00 a copy. A court ruled that Amateur and CB license fees were illegal, then made the FCC give back all the fees it had collected. Ever-tightening federal budgets forced the Commission into serious cost-cutting, and one of the more positive results is the very successful Volunteer Examination program that's made access to taking Amateur exams easier than it's ever been before.

The very face of Amateur Radio has changed drastically in the past 13 years. In 1974 vhf/uhf fm, though popular, supported fewer than a thousand FCC-licensed Amateur repeaters; ARRL's latest repeater directory lists well over 10,000! The separate "WR" repeater license, along with secondary station licenses and new club and RACES licenses, all fell victim to FCC "deregulation." Computer technology became an integral part of Amateur Radio and is, indeed, an integral part of many Amateur radios. New computer-based hard copy/data communications modes such as AMTOR and Packet, unheard of just a few years ago, have far eclipsed their predecessor, RTTY, in their number of enthusiastic users.

On the Amateur bands, DXers, vhf/uhf "weak signal" buffs, traffic handlers, and rag chewers alike have enjoyed the benefits and suffered through the depths of a little over one complete sunspot cycle. For hf DXers, there's been the first legitimate activity from BY and YI for many decades, along with a fair number of brand-new "new ones" to keep Honor Roll members and aspirants close to their shacks. We've reported on earthquakes, hurricanes, floods, tornadoes, blizzards, forest fires, and even a volcanic eruption during the past 13 years, and of the contributions of the countless thousands of Amateurs who've responded with skills and equipment to help alleviate the resulting human suffering.

It has indeed been an exciting 13 years for Amateur Radio — though one must wonder if it's really been all that much more so than any similar period of time in the 1920s, 1930s, or post-war years. After all, exciting things have always happened in Amateur Radio, and will, we trust, continue to happen. It's been fun being part of it, as both a participant and a reporter.

Thus it is, with a real feeling of regret, I'm retiring as editor of *Presstop*. After 13 years on the reporter's beat for *ham radio*, a new opportunity outside of Amateur Radio has forced me to relinquish that aspect of my Amateur Radio involvement. That doesn't mean I'm giving up either Ham Radio or *ham radio* — you'll still find me on the air enjoying the former and, when time and my New Hampshire colleagues permit, on the pages of the latter.

73,
Joe Schroeder, W9JUV

presstop de W9JUV

CHALLENGING IDEAS FOR AMATEUR RADIO'S FUTURE, particularly with respect to the concept of "personal communications," are contained in the recently released FCC Office of Plans and Policy Working Paper No. 20. With respect to the Amateur Service, author James McNally, WB3APV, notes that Amateur Radio is in a slow growth phase that could continue for many years under present limitations, then suggests such revolutionary ideas as relaxing the limits on "permissible communications," permitting an Amateur's immediate family limited use of his or her station on portions of the VHF/UHF bands, and/or creating a combined Amateur/personal radio service on parts of the 33- or 23-cm bands. Also discussed at length is the GMRS service, which he also urges be greatly expanded and made more flexible.

This Fascinating And Very Thought-Provoking Study should be required reading for anyone who is concerned with the future direction of the Amateur Radio Service.

NOVICE "ENHANCEMENT" PROBABLY WON'T APPEAR IN ANYONE'S XMAS STOCKING this year, despite optimistic predictions from several sources. Though the proposal to give many new privileges to Novices (and Techs) is enjoying broad Amateur support, the usual timetable for seeing a significant proposal through FCC procedures makes such a short turnaround seem unlikely. A more realistic estimate would put Commission action into early 1987.

FCC HAS PROPOSED 40-METER NOVICE BAND EXPANSION TO INCLUDE 7050-7075 kHz in Alaska and U.S. Carribbean and Pacific islands, due to interference in the 7100-7150 kHz Novice band. Comments on the NPRM, PR Docket 86-397, are due December 22; Reply Comments by January 21.

Automatic Control Of VHF/UHF Packet Stations using ARRL AX.25 protocol has been approved by the FCC, effective November 24; automatic HF digital operation wasn't decided.

The Effective Date For VECs To Be Responsible For Maintaining the Amateur examination question pool was still pending at presstime. It should, however, be before year's end.

The Comment Due Date For FCC's Automatic Transmitter Identification System proposal (see October Presstop) has been extended to January 19, 1987; Reply Comments to February 16.

ALLEGATIONS OF FRAUD IN AMATEUR EXAM ADMINISTRATION ARE CASTING A PALL over the VEC program. Serious questions have been raised about certain VE teams in Texas, Puerto Rico, North Carolina, and southern California, and all are believed under FCC investigation. In at least one case there's also reported to be a "repeater war" between some of those involved that may be related to the accusations, but there also appears to be grounds for belief that at least some limited "selling" of Amateur licenses has taken place.

W5YI Reports Two Newly Licensed Amateurs Examined by one of his VE teams were called in for reexamination by a DeVry VE team at FCC request, and that he's removed the accreditation of another of his VEs in California. Fortunately, in terms of the VEC program's size, the improprieties involve only a very small minority of the participants.

THERE'S SOME ENCOURAGEMENT IN THE FCC'S LATEST AMATEUR GROWTH FIGURES, which show another all-time high of 421,082 U.S. Amateurs at September's end, though that's only a 2 percent increase over a year ago. What's more encouraging is the number of new Amateurs licensed during the past 12 months — 20,979, up 20.75 percent over the preceding year. In the same period dropouts decreased by 15.13 percent, to 12,484. By class, Extras are up by 7.38 percent, Advanced 4.73, Tech and Novice 3.6 each, while the only decrease was for General, down 0.5 percent. 22,228 Amateurs upgraded during the year, a 10 per cent increase.

"ARCHIE'S HAM RADIO ADVENTURE," THE INDUSTRY-GROUP-INSPIRED comic book designed to pique youngsters' curiosity about Amateur Radio, is now available in quantity. The story line in this very well done 36-page booklet manages to include HF and VHF fm, DX, OSCAR, autopatch and even packet radio while maintaining an entertaining and coherent story line. Educators and others in a position to place copies in the hands of youngsters in the targeted 10-14 year old age group should contact Viki Armentano at the ARRL.

WESTLINK REPORT'S FIRST "YOUNG HAM OF THE YEAR" IS SHAWN WAKEFIELD, WK5P, a 16-year old Eagle Scout from Bartlesville, Oklahoma. Not satisfied with getting five of his fellow Scouts into Amateur Radio, Shawn led efforts to set up Radio Explorer Post 103 and also organized a radio club in his high school, where he's an honor student. Active in RACES, he also set up and runs a 7274-kHz railroad buffs' net on alternate Tuesdays at 0000z.

Shawn Was The Judges' Unanimous Choice, despite strong competition from several other nominees. He was to receive the award, which includes a complete Yaesu station, at the Ham/West Convention banquet in Las Vegas November 8.

PRESSTOP EDITOR JOE SCHROEDER, W9JUV, IS RETIRING as of this issue, and ham radio's editors have decided not to continue the monthly Amateur Radio news column without him. Joe's just purchased a small gun collecting book publishing firm, and expects his new responsibilities there to keep him fully occupied. He will, however, remain on the ham radio masthead to contribute editorials and an occasional feature article.



comments

buying topo maps

Dear HR:

Here's some updated information regarding the purchase of topographic maps from the United States Geological Survey.

First, the headquarters of the United States Geological Survey has been moved to Reston, Virginia 22092. All mail orders, however, should be addressed to the Distribution Branch, United States Geological Survey, Box 25286, Federal Center Building 41, Denver, Colorado 80225.

For a copy of the USGS index to topographic maps, call 703 648-5830. Request the index for a specific geographic region — for example, Maryland, Delaware and the District of Columbia, which are grouped together. Each regional index lists local retailers who carry USGS maps as well as a variety of special maps, including state maps, satellite image maps, and maps showing national parks, rainfall, population distribution, and similar information.

I hope this information will be useful to those seeking that ideal site on which to place their "Giant Killer Contest Stations."

Dan Szymanski, K3SKE
Frederick, Maryland 21701

cheers from down under

Dear HR:

Received July's *ham radio* yesterday and was fascinated by K2RR's "Reflections" ("The Phone Call," page 4).

The problems discussed have also been part of an investigation by the Redcliffe Radio Club. . . . We also discussed the problem a few weeks ago and came to a very interesting conclusion, which is probably worth following up in the USA as well.

Attracting youngsters to Amateur Radio is pretty difficult, what with all those distractions — including computers. It's easier, especially if one has access to a modem, to use the 600-ohm twisted pair. Then too, Australian school kids are generally not interested in studying extra-curricular subjects because they don't count as academic credits. Why work hard for something that doesn't count?

A far greater source of new Amateurs would appear to be those approaching retirement age, who are trying to find something to do besides fishing, hunting, or annoying the "missus." This last suggestion has been based on past experience within the Club. We found that most of the adult students persevered and finally passed the Novice exam and within a year upgraded to either Limited License (your Technician License) or General License, while out of ten young students (under 20 years) only three or four made the grade; all the rest all dropped out.

Due to the nature of the Amateur organization in Australia — where, because of the cost, most work is done by volunteers, and not much can be done by either the Federal Council of the WIA or the State Councils — much depends on the Clubs. And of those, only a handful are trying their hardest to recruit more Hams.

There is one danger [in recruiting new hams], however. A club specializing in classes for newcomers may tend to forget those who passed the exami*nations simply because all the* volunteers are involved in training or other work to keep the Club afloat.

I would like to hear of US Clubs who have been able to keep licensed Amateurs interested enough to remain active in the Club.

Thanks for a magnificent magazine — and 73 from Down Under.

John Aarsse, VK4QA
Nambour, Australia

short circuits

Digi-Key address

In the October, 1986 article, "Get on SSTV—with the C-64," the address of Digi-Key, Inc., was stated incorrectly (**fig. 1**, page 45). The correct address is Digi-Key, Inc., P.O. Box 677, Thief River Falls, Minnesota 56701-9988.

low-noise UHF/VCO

In part 1 of my article, "Low-Noise Phase-Locked UHF VCO" (July, 1986, page 33) the noise floor in fig. 3 was shown as −142 dBc/Hz for N = 1. To label the curve shown with N = 1, however, is ambiguous, since the phase detector compares 1152 MHz with the 32nd harmonic of the reference crystal oscillator.

Reference oscillator noise increases by 20 log N where N in this case is 32. Therefore, the noise floor of the 1152 MHz VCO should be moved to about −130 dBc/Hz, if the effect of reference multiplication and noise in the loop is taken into consideration. This puts the noise floor corner at about 20 kHz, with noise rising at 20 dB per decade to around −100 dBc/Hz at 200 Hz. These numbers compare favorably with those of commercial synthesizers currently available.

The UHF VCO PLL discussed in the article is a "direct" type, meaning that the phase detector corrects VCO phase at 1152 MHz, not some submultiple of 1152. It is difficult to achieve better phase noise performance by any other means. — WA9HUV

vhf/uhf world

The schematic shown in **fig. 3** of W1JR's November, 1986 column, "VHF/UHF World: Broadband Amplifiers in Receiver Design" (page 91), should be positioned in **fig. 4**. Likewise, the schematic shown in **fig. 4** should be positioned in **fig. 3**. Only the art should be switched; captions are correctly placed.

communicating on 474,083 GHz

Explore a new frontier with your own lightwave station

Steve J. Noll, WA6EJO, TiC Scientific, 1288 Winford Avenue, Ventura, California 93004

Although we might wish for bigger slices of the HF or VHF spectrum, Amateur Radio's frequency allocations extend over 23,600 MHz.

In assigning frequencies, the FCC didn't terminate our allocations at some obscure, arbitrarily chosen microwave band. Instead, they specifically gave us everything above 300 GHz. And although Amateur communication above the 24-GHz band isn't yet very practical, there is a narrow segment where it's really quite easy to work: that is, of course, *light*.

background

I made some of the first Amateur two-way laser contacts during ARRL VHF contests. At that time (June, 1979) the ARRL didn't recognize contacts above the normal microwave bands, even though the FCC regulations did. These contacts, however, were rejected.

I figured that the League would be more ready to accept lightwave contacts if specific guidelines for such contacts were available. With the encouragement of Louis Ancioux, WB6NMT, I began drafting recommendations. In mid-1980, I sent the ARRL Contest Advisory Committee several suggestions for rules.

This is what was ultimately adopted:

Above 300 GHz, contacts are permitted for contest credit only between licensed Amateurs of Technician class or higher, using coherent radiation on transmission (e.g., lasers) and employing at least one stage of electronic detection on receive. . .

The rule also state that while no minimum distance is specified for contacts, equipment should be capable of real communications (i.e., able to communicate over at least a mile). This was no problem at light frequencies; my first laser contacts were over a 15-mile path!

It should go without saying that all equipment used for contest points or records should be Amateur-owned, if not also Amateur-built. After all, there would be no technical challenge or achievement in borrowing a NASA dish to make a 432-MHz moonbounce contact.

what is light?

Light is a part of the electromagnetic spectrum. Visible light starts at around 400 Terahertz; 1 THz = 10^6 MHz. The low-frequency, long-wavelength end of the visible spectrum appears to us as the color red. The high-frequency, short-wavelength end is violet. Green, in the center of the visible spectrum, is the wavelength to which our eyes are most sensitive.

Like "regular" radio waves, light has a wave nature with an electric field and magnetic field component. It also has an associated frequency (which is only rarely referred to) and wavelength (the more customary measurement).

A variety of length measurement units are used to describe light wavelength. Green light, for example, corresponds to approximately 5500 Å (Angstrom Units where $1Å = 1 \times 10^{-10}$ meters), 550 nanometers, 0.55 microns, or a frequency of 545 THz. But even though it follows many of the same rules as rf, light isn't just a radio wave of extremely high frequency. What sets light apart from lower frequency emissions is its particulate nature. Light exists in discrete units called photons and can be made to travel in a very narrow beam with very little divergence or spreading. It's wonderful for point-to-point communications.

Frequency multiplication via a non-linear varactor is a phenomenon well known to the UHF enthusiast. 432 MHz is conveniently reached by tripling 144 MHz. What is both amazing and not commonly known, however, is that light frequencies can also be multi-

plied in a special non-linear crystal! But light can not only be multiplied in frequency; it can be mixed. It's possible to direct two laser beams into a special crystal and produce a third beam of a different frequency as a result of the mix. It should therefore be possible to build some sort of an optical superheterodyne receiver. It's also possible to build a light "amplifier," though the necessary crystals and lasers are very specialized and expensive.

ceiver. It's also possible to build a light "amplifier," though the necessary crystals and lasers are very specialized and expensive.

light transmitters

The laser was conceived in 1958 and built in 1960. "Laser" is an acronym for Light Amplification by Stimulated Emission of Radiation.

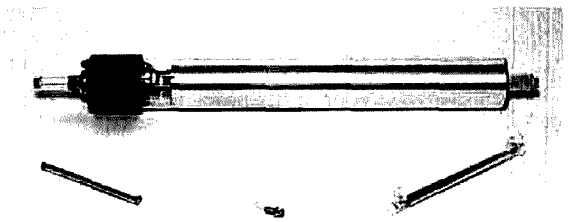


Photo A. Four types of lasers: from top (clockwise), gas laser (helium neon); solid-state laser (ruby rod); semi-conductor laser (infrared pulsed diode type); and dye laser (dye cell filled with organic laser dye Rhodamine B.)

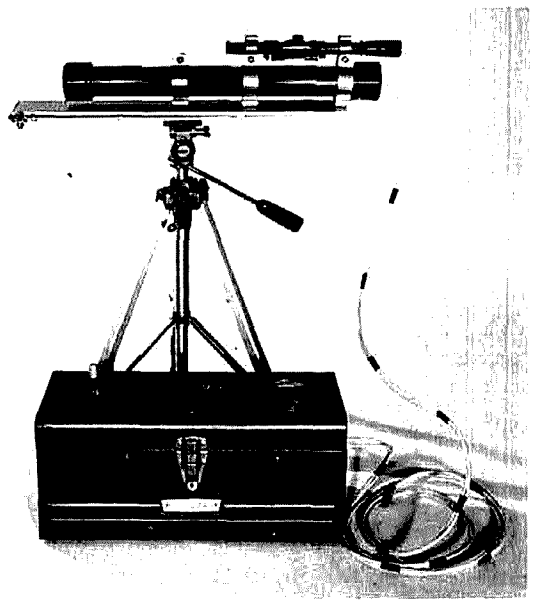


Photo B. This Helium Neon laser uses the power supply circuit shown in fig. 7. *Power supply and* battery are housed in a tool box.

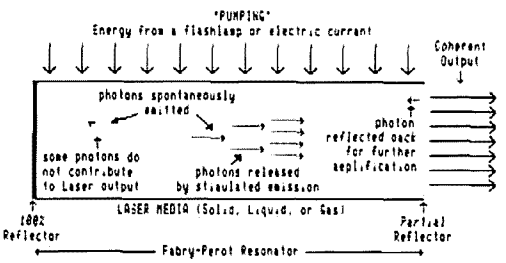


fig. 1. Basic laser operation.

This is how lasers work (**fig. 1**). A gas, solid, or liquid is contained in a cavity. Two mirrors, one at each end of the cavity, are positioned to face each other. Energy is transferred to the atoms inside the cavity that do the actual "lasing"; depending on the kind of laser, the energy can come from a flashlamp, an electric current, a chemical reaction, other energized atoms or molecules, or another laser.

The goal is to raise a lot of the electrons of the lasing atoms to a certain higher energy level. At this level, the atoms are unstable and won't stay there long. When the electrons "fall down" to a lower energy level, they give up this excess energy, which is released in the form of photons. This release of photons — random in time and direction — is called *spontaneous emission*.

If one of these photons strikes an atom that hasn't yet generated its photon, the impinging photon stimulates the atom to produce another photon identical to the first. The result is two photons of the exact same frequency and phase, traveling in the same direction. If everything is just right, this process continues through the lasing medium.

Some of the photons bounce between the mirrors, traveling through the medium, stimulating more coherent photons to be generated. These, then, also bounce between the mirrors, continuing the process. One mirror is partially transparent to let some of the energy escape. Thus is born the laser "beam."

Although most people have heard of the ruby laser and the Helium Neon (HeNe) laser that's commonly used in supermarket scanners, a great number of other materials have been used to make lasers. Most can be categorized as *gas, solid-state, liquid,* or *semiconductor* (**Photo A**).

gas lasers

Output wavelengths range from infrared to ultraviolet. Some are tunable. Power outputs vary from extremely weak to strong enough to burn through steel plate. Outputs vary from a few short pulses per second to a continuous wave. Sizes vary from as small

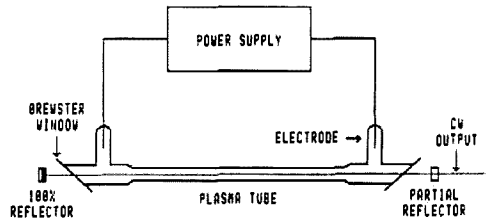


fig. 2. Basic gas laser.

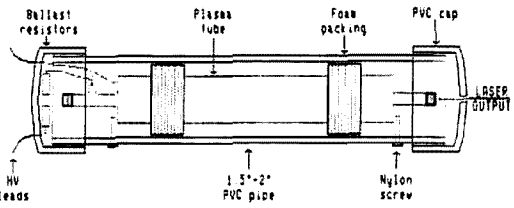


fig. 3. HeNe laser plasma tube packaged in PVC pipe.

as a pencil tip to room-size. And cost ranges from a few dollars (for surplus semiconductor lasers) to hundreds of thousands of dollars.

Known for its high-power infrared output, the carbon dioxide laser is capable of burning through metal. The Argon laser can generate several watts of green and blue light. The Nitrogen laser has an ultraviolet output and so much gain that cavity mirrors are not needed.

Most gas lasers (**fig. 2**) are electrically excited by applying high voltage to electrodes at the ends of the cavity, also known as the "plasma tube" (**fig. 3**). Sometimes windows seal the ends of the tube to let the light through to strike external cavity mirrors. Often there are no windows, and the mirrors are attached directly to the ends of the plasma tube. Gas lasers usually have lifetimes expressed as a few thousand hours, after which time some kinds may be cleaned and refilled with a fresh charge of gas.

solid-state lasers

In this application, "solid-state" doesn't refer to semiconductors, but instead to lasers made of crystal or glass. One of the better known of these is the ruby laser, which consists of an Aluminum oxide "host" with a tiny bit of Chromium mixed in. The laser may take the appearance of a pink ruby rod measuring 1/4-inch diameter and a few inches long, positioned close to a long flashlamp tube. Mirrors may be external or deposited directly on the rod's ends. The ruby laser puts out short, high-power pulses of red light.

The ruby laser may be the most well-known solid-state laser, but the Neodymium YAG laser finds the most use. Capable of continuous output in the 100-watt range, its output is invisible infrared. As with the ruby, the Nd/YAG laser is optically pumped with flashlamps or continuous arc lamps. It is used for cutting and drilling, trimming resistors, and surgery.

liquid lasers

Liquid lasers are often called "dye lasers" because the active material is usually an organic dye dissolved in a liquid solvent. The liquid laser's claim to fame is that it is tunable over a wide frequency range.

In the simplest form of dye laser, the dye is held in a length of glass tubing, its ends sealed with windows. The external mirrors form the lasing cavity and a flashlamp serves as the pump source. CW-output dye lasers are available. Power levels are generally under 5 watts.

semiconductor lasers

Electro-optically similar to Light Emitting Diodes, semiconductor lasers are often called *diode lasers* or *injection lasers*. Output is typically in the infrared,

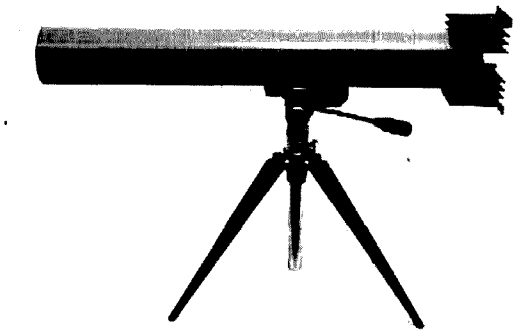


Photo C. This helium Neon laser is housed in a 2½-inch diameter PVC pipe, which also contains a commercial 12-volt inverter power supply.

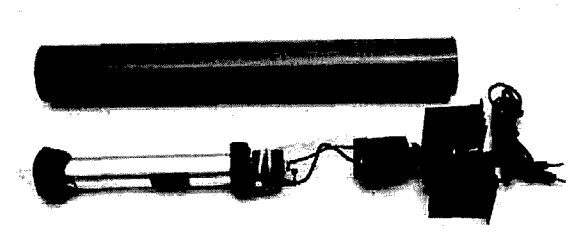


Photo D. Components of a portable HeNe laser.

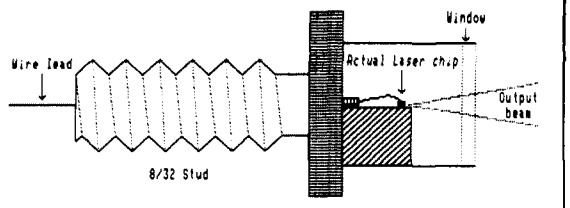


fig. 4. Diode laser in a typical stud package.

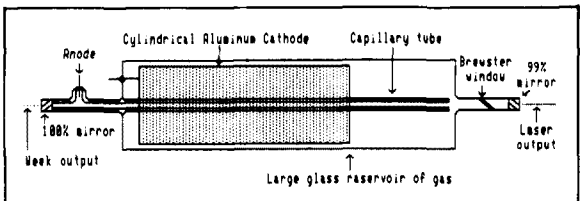


fig. 5. Modern HeNe laser tube.

measuring a few milliwatts CW or several watts pulsed.

A common package is a stud-type with a window on top to let the radiation out (**fig. 4**). Lasing action is achieved by running current through the tiny diode "chip." Mirrors are formed directly on the chip by cleaving opposing ends of the chip square.

This kind of laser doesn't produce a beam as narrow as most other lasers. Pulsed infrared diode lasers are presently on the surplus market — often for less than $25.00. We expect to see many surplus units in the future because of the growing fiber optic communication and compact audio disk markets.

making the choice

It's not difficult to narrow down the wide choice of lasers to a few suitable for our use. Many lasers are quite expensive, costing several thousand dollars or more. Some are quite bulky because of large power supplies or cooling apparatus. Some produce short bursts of light only at a very low pulse rate. A few are downright dangerous, emanating invisible beams capable of burning through almost any known material.

For communications use we need a laser with the following properties:

- It must be inexpensive, or at least so once it reaches the surplus market.
- It must offer reasonable lifetime — at least a thousand hours between refurbishment or replacement — consistent with its cost.
- It should be small, portable, and not consume a lot of power.
- It must be reasonably safe to operate and work with.
- Although an infrared output that matches readily available infrared detectors would be usable, a visible beam is preferred.
- A continuous wave output is most desirable. A pulsed output with a very high repetition rate could be used.
- There should be a way to modulate the output and then electronically detect the result.

Solid-state lasers are generally quite expensive, infrared, pulsed, and not often available on the surplus market. The liquid lasers are also rather expensive and not available as surplus. They usually have a visible beam, but tend to be bulky.

Semiconductor lasers present some possibilities. The smallest of all lasers, they're inexpensive, readily available, and offer long life. They can be battery-powered. Power output is generally low enough to be considered safe.

Their biggest disadvantage is that the affordable diode lasers put out only infrared radiation; visible light is much easier to work with. Luckily, infrared detectors are plentiful.

The beam from a diode laser is broader than most others and requires shaping by additional lenses or mirrors to make it travel a long distance without too much divergence. The surplus units are usually pulsed only, but the repetition rate is high enough to be useful.

There are many choices of gas lasers, but only a few meet our criteria. The CO_2, Argon, Krypton, and metal vapor units are typically expensive, not available as surplus, large, short-lived, energy-hungry, and dangerously powerful.

One candidate looks especially good, however: the HeNe laser, known as the "light bulb" of the laser industry (**fig. 5, Photo B**). Luckily, the laser that's one of the most suitable for Amateur use is also one of the most common on the surplus market.

Actually, what one usually finds is the plasma tubes, not complete lasers. The choice of a power supply and packaging are left up to the buyer.

The HeNe laser is fairly affordable — under $300 for a small, brand-new, complete unit. Surplus plasma tubes, with integral mirrors but without a power supply, often sell for under $100, and sometimes as little as $30. The HeNe's lifetime is many thousands of hours, i.e. a few years. Its size is reasonable, and power consumption is under 30 watts for the smaller units. The plasma tube itself generally requires up to 2000 volts dc and up to 10 mA. A starting voltage of 10 kV may be required. The high voltage requirement of the plasma tube is perhaps the greatest hazard associated with the HeNe laser.

The common off-the-shelf HeNe lasers have power outputs rated from 0.5 mW to 5 mW. The beam power of the under 1-mW models is high enough to cause blindness. The higher power units will definite-

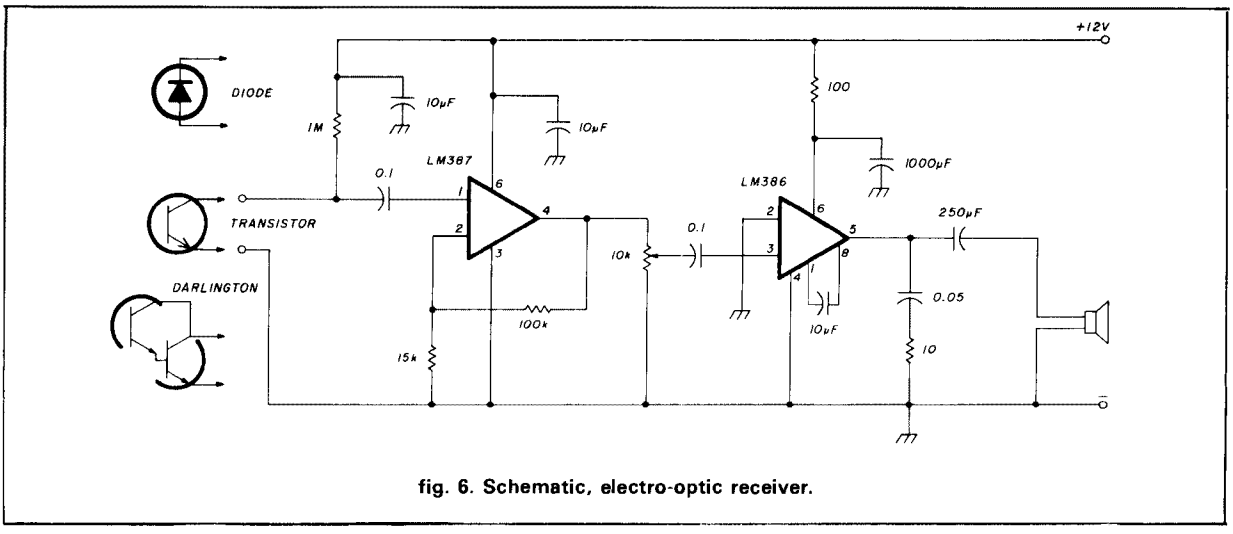


fig. 6. Schematic, electro-optic receiver.

ly require reasonable care in handling, but they're still not capable of burning through things.

The HeNe's beam is bright red, continuous, and very narrow — perfect for Amateur uses. (It's the one I've chosen to work with.) Beam diameters are typically 0.5 to 1 mm and divergence is typically 1 to 2 milliradians, which is extremely tight. Plasma tube sizes are usually 6 to 15 inches long and 1 to 1-1/2 inches in diameter. *Don't buy from anyone who won't guarantee that the laser works. If the plasma tube is dead, there's nothing you can do to fix it.*

The only other major item necessary is a power supply, which, of course, can be homebrewed. There are some nice commercial modules that run from 12 volts dc and supply the starting voltage as well as the high voltage running current.

If you're starting with a plasma tube, a good way to package it is inside a length of PVC pipe. An excellent insulator, PVC pipe is easy to work with and inexpensive. A diameter of 1-1/2 or 2 inches is about the right size (see **Photos C, D**).

HeNe lasers require a ballast resistor which can be mounted in the non-output end of the pipe, inside of an end cap. You might be able to fit the whole power supply in one end of the pipe.

modulating lasers

There are two ways to impress information on a laser beam. One is to modulate the laser itself — that is, to make it generate light in varying amounts. The second way is to insert something the path of the unmodulated beam.

An obvious way to modulate the generation of laser light would be to modulate the power source. In the case of the HeNe laser, this would mean modulating the nominal 2-kV dc power supply.

Unfortunately, this approach doesn't work as well as we'd like. The HeNe laser — sort of a glorified Neon sign — is built around an electrical gas discharge. Perhaps you're familiar with the operation of the common Neon pilot light. The NE51, for example, has a threshold beneath which it will not light. If much more than the rated voltage is applied, however, the lamp will be ruined. It doesn't like its power supply voltage or current to vary very much at all.

Modulation can be achieved via the power supply, but the percentage of modulation will not be high. Even some commercial modulated HeNe lasers, which are specifically made to be modulated, offer only 15 percent modulation.

Power supply modulation does offer some advantages. It won't, for example, distort the laser beam shape. This is very important for long distance atmospheric communications where an extremely narrow collimated beam is necessary.

One form of power supply modulation is to place a modulation transformer in series with one of the high-voltage lines leading to the laser plasma tube. I've tried this using a Collins modulation transformer. A loud signal was picked up by the 931 photomultiplier receiver, although the modulation level was less than 1 percent. Attempts to increase the modulation level simply extinguished the plasma; apparently the modulation causes the high voltage to drop below the plasma threshold. A complete receiver is shown in **fig. 6**; a laser power supply schematic is shown in **fig. 7**.

Modulation in the beam path avoids the previously-mentioned power supply problems. 100 percent modulation *is* achievable. There are a variety of objects that can be used as modulators.

A "chopper" usually takes the form of a disk with notches or holes cut into it (**Photo E**). Powered by a small electric motor, the spinning disk cuts the beam of light. While this may seem like a crude approach, such devices are serious laser tools and can often be seen advertised in the laser-related trade magazines.

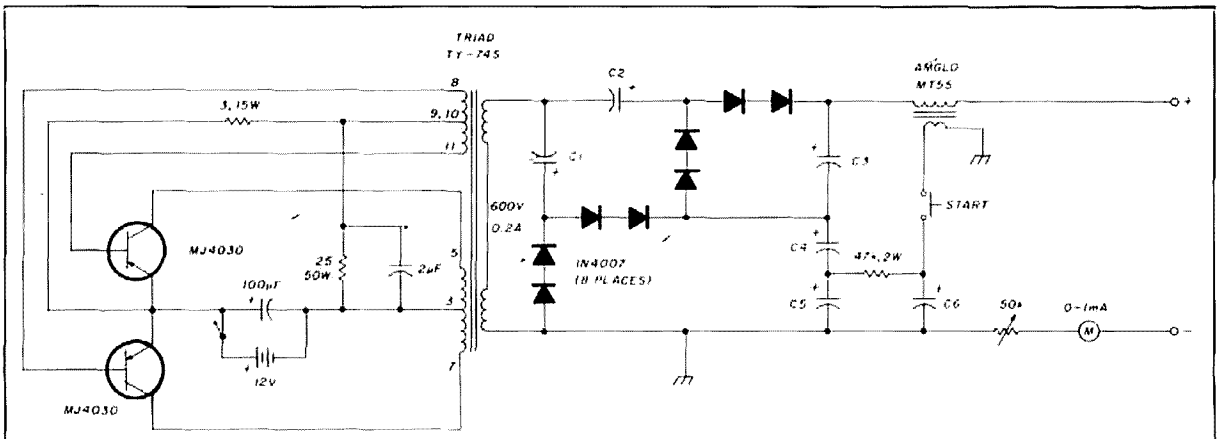


fig. 7. 12-volt dc inverter HeNe laser power supply.

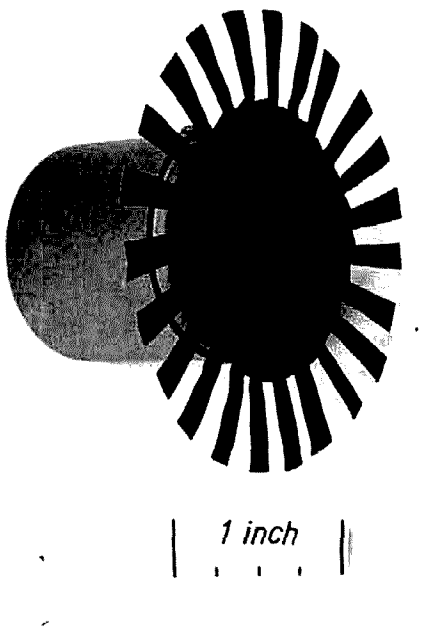


Photo E. Optical chopper made from a small dc motor and tin can lid.

A chopper certainly provides 100 percent modulation. The beam is either on or off, and is not distorted or deflected. With a reasonable motor speed and enough holes or slots in the disk, the beam will be interrupted at an audio rate. The audio-modulated beam then can be interrupted to produce type A2 modulation, modulated CW. The Morse code interruption of the beam can be done with a solenoid activated by a telegraph key or simply by hand.

I used this simple chopper method of modulation for my first contacts. The chopper disk was made from a 3-inch diameter tin can lid. I cut two dozen slots in the perimeter with tin snips, then soldered the center of the lid to a toy electric motor shaft. I clamped the assembly to the side of the laser housing and powered it with a 9-volt transistor radio battery. The resulting audio tone-modulated beam was interrupted by hand to achieve the final Morse code modulation. This crude scheme worked quite well and resulted in a very loud signal over a 15-mile path!

Several kinds of electro-optical modulators use special glasses, crystals, or liquid cells. When high voltage, rf, or a magnetic field are applied, the polarization of light passing through the modulator is rotated. Unfortunately, prices for these devices run from several hundred to tens of thousands of dollars.

Liquid crystal displays are light modulators that have become part of our everyday life. Wrist watches, portable computers, and electronic gear of all kinds use LCDs as light valves to display information. It's possible to modulate light by beaming it through a transmissive type LCD or a reflective LCD with the reflector removed (**Photo F**). The main drawback is speed; the spec sheet for one common display lists the turn-on time as 100 milliseconds and the turn-off time as 200 milliseconds — rather slow for audio frequency modulation. Faster LCDs may become available in the future as the need for rapid graphics displays grows.

the receiver

Unlike trying to modulate light, receiving modulated light presents no serious problems.

Like a simple radio receiver, a light receiver has an antenna for gathering energy, a "front end" filter for selecting the wavelength of interest, a detector for converting the photon energy to electrical energy, and miscellaneous electronics for amplification of the final signal (**fig. 8**).

the "antenna"

As with radio receivers, the antenna of a light com-

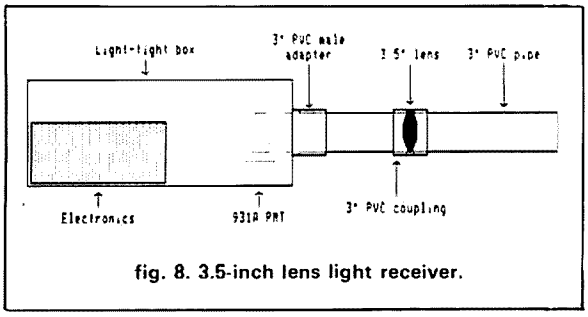


fig. 8. 3.5-inch lens light receiver.

munications receiver presents the best opportunities for system improvements. ("The bigger the better" applies here also.) The purpose of the light antenna is to gather as much light from the desired source as possible, concentrate this light onto the detector, and reject light coming from unwanted sources — i.e., the sun, moon, street lights, and other sources of ambient light. Lenses or curved mirrors mounted in an appropriate tube or holder make a complete antenna system. Obviously, what we need is something like a telescope, though we needn't bother trying to obtain an actual detailed image of the light source. We need merely to gather as much light from the source as possible.

Large-diameter lenses make good antennas. Because of weight and cost, 3 to 4 inches in diameter is about the largest size in which glass lenses are commonly found. The diameter of surplus lenses under $20 usually doesn't exceed 3 inches. (Mirrors take over from there.) One source of large-diameter lenses is stationery or office supply stores that sell magnifying lenses for reading small print. These lenses may be made of plastic instead of glass and will require a little more care in handling to prevent scratches. Also, watch flea markets and other sources of used equipment for Luxo-type magnifying work lights.

Fresnel lenses are flat lenses, often plastic and less than 1/8 inch thick. One side is smooth while concentric ridges are molded onto the other side. Fresnel lenses won't form much of an image, but they will gather a very large amount of light for their cost and weight. I use a 10-inch diameter Fresnel lens mounted in a metal duct pipe to serve as an antenna for one of my portable photomultiplier receivers.

Parabolic and spherical curved mirrors make excellent light antennas, as witnessed by their heavy use in optical astronomy. Swap meets and garage sales may be good sources for old reflector telescopes built around such mirrors. A small telescope with a photosensor mounted in place of the eyepiece would make a great light communications receiver, but a large Fresnel lens would probably gather much more light. Curved sheet-glass mirrors and spun aluminum parabolic reflectors measuring 1 to 2 feet in diameter may work well in spite of their inaccuracies at optical wavelengths.

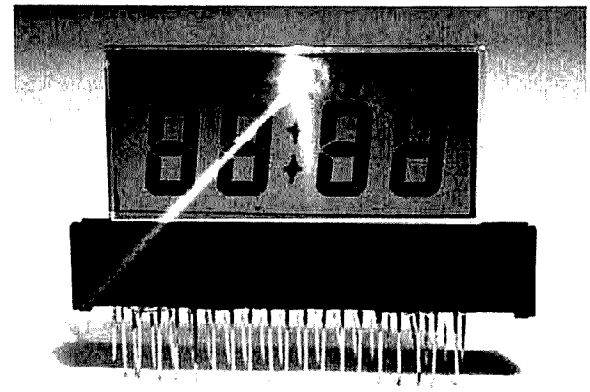


Photo F. A common LCD is used as a laser light modulator. Laser beam is seen entering at bottom left.



Photo G. Portable laser communications receiver uses a 10-inch diameter Fresnel lens for the "antenna" and 931A photomultiplier tube as the detector.

If only nighttime communication is anticipated, optical bandpass filtering can be dispensed with. But daylight communications is a different story; sunlight can swamp the photodetector, limiting your communications range. A very narrow bandpass filter is required. Luckily, dielectric filters, sometimes called interference filters, fit the bill. These filters are made by depositing multiple thin layers of chemicals on a glass substrate. (Edmund Scientific' sells these filters for about $20 each.) The specifications for their HeNe laser filter state a half-power bandwidth of only 1.7 percent! The drawback is that the filter insertion loss is 55 percent, which isn't too bad considering the bandwidth.

detectors

As it is with lasers, there are a large number of different photo detectors available. There are photodiodes,

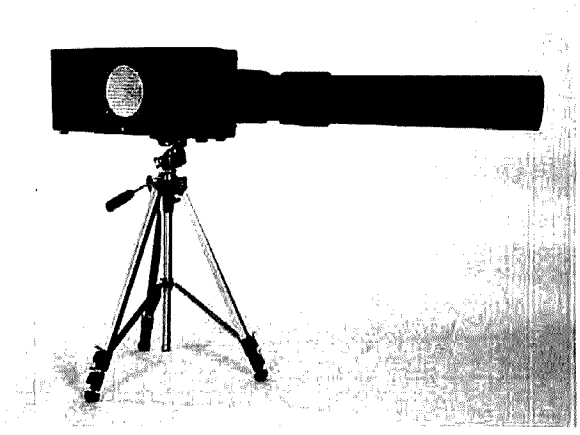


Photo H. Photomultiplier tube communications receiver. A 3-inch diameter lens is housed in the long PVC pipe.

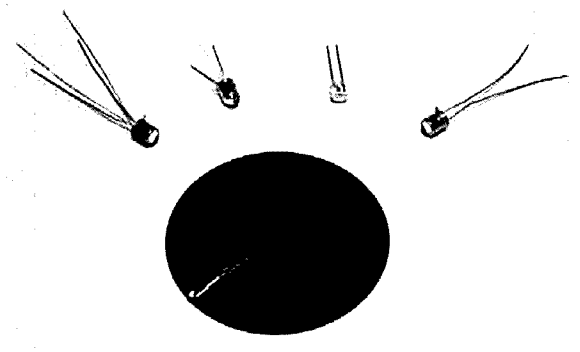


Photo I. Solid-state photodetectors. Left to right: photo FET, bipolar photo transistor, photo Darlington transistor, photo diode. Bottom: photovoltaic (solar) cell.

phototubes, photomultiplier tubes, phototransistors, photoFETs, photovoltaic cells (solar cells), photoresistive cells, avalanche photodiodes, PIN photodiodes, photo Darlington transistors, and even photo traveling wave tubes!

Next to cost, there are three other important areas of concern. The first is *spectral response*. Does the device function well at the wavelength of light we are dealing with? Presumably we will be mainly concerned with the red output of a HeNe laser. Many kinds of detectors do not perform well at this wavelength. The second is *sensitivity*; the light sensitivities of the different kinds of sensors vary by at least a factor of 1000. *Speed*, the third, is mainly a concern with short pulse width applications such as receiving light generated by a diode laser. It is not a concern with audio or mcw continuous output lasers (HeNe).

Solar cells are not considered to be especially good for light communications. Although they are inexpensive and fast enough for audio detection and their "capture area" is very large, this is not an advantage in situations where lenses or mirrors are used to gather light. They are also easily overloaded by sunlight.

Photoresistors are inexpensive and have good sensitivity at red light, but are too slow for detecting modulation.

There are several kinds of photodiodes. PIN photodiodes and Avalanche photodiodes are the best kinds for communications. The Avalanche photodiode is the most sensitive and the fastest. Unfortunately, it is also the most expensive and is generally not found on the surplus market.

The PIN photodiode is an excellent choice. It has a response time on the order of one nanosecond and its sensitivity at red light is good. The Motorola MRD500 and MRD510 diodes sell for about $3.50. The Motorola Optoelectronics data book describes these two diodes specifically as being suitable for laser detection and light demodulation.

In the category of phototransistors, there are bipolar phototransistors, photo Darlington transistors, and

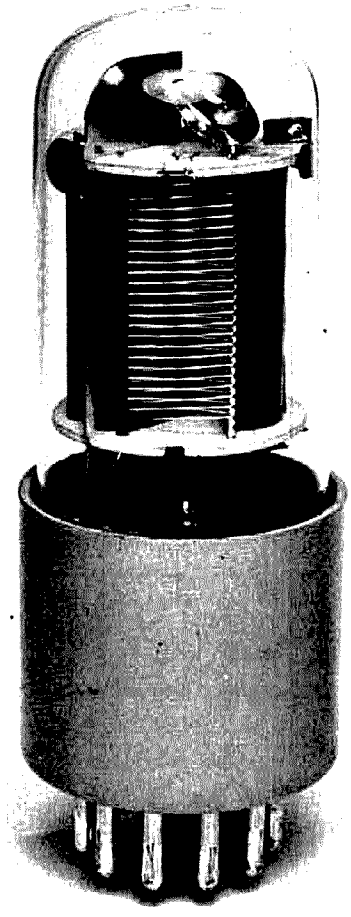


Photo J. 931A photomultiplier tube.

photoFET (**Photo I**). Phototransistors are more sensitive than photodiodes since they have gain. This gain is achieved at the price of speed, although they are still fast enough for audio detection.

Bipolar phototransistors are cheap and commonly available. A search of electronic parts catalogs yielded no fewer than 15 different devices available with prices ranging from 30 cents to less than $3. The response time is around 2 to 20 microseconds and sensitivity to red light is good. They can be saturated by too much light and care should be taken to prevent this by optical shielding and filtering.

The photo Darlington transistor contains a light-sensitive transistor as in the bipolar phototransistor and an additional transistor that amplifies the output of the first one. These devices are very light-sensitive. Speed is correspondingly slower, in the 10 to 100 microsecond range. The Motorola MRD360 photo Darlington is one of the best units and sells for under $5.

Photomultiplier tubes are bulky and fragile, require high-voltage power supplies, cost more than most of the common solid-state devices, and can be hard to find. Their spectral response is usually not the best at red. So why use them? Gain. Lots of gain. Gain into the *millions*. The gain is developed inside the tube and without excessive noise. Rise times of 3 nanoseconds are typical. Photomultiplier tubes can detect extremely low levels of light and are linear over wide ranges of light intensity. The popular 931A (**Photo J**) is linear within 3 percent over a change in light intensity of 10,000,000.

The Avalanche photodiode is the closest solid-state equivalent, but it is even more expensive, extremely temperature-sensitive, and requires a highly regulated high-voltage power supply. If Avalanche photodiodes were ever to become low-priced surplus items, they might then offer some advantage over the photomultiplier. But for now, the "hottest" optical front end is the photomultiplier tube.

The common photomultiplier tubes look similar to a glass envelope octal base radio tube. They typically have eleven or more base pins, depending on the number of stages. There is no filament. The 931A and 7117 are two of the more common tubes. The peak sensitivity is at 4000 Å (violet) and is unfortunately less than 10 percent at red wavelengths. With a 1000-volt supply, the current amplification for the 931A is 800,000.

These tubes were once used in automatic headlamp dimmers, so one source of them would be auto wrecking yards. The Allied Electronics[2] catalog lists the 931A at $34 (new) and Fair Radio[3] sells used units for under $10.

propagation

Obviously propagation is limited to line-of-sight. The range depends on laser power, wavelength, and divergence. The wavelengths we're dealing with travel very well through the atmosphere. The beam divergence is typically one milliradian, which means for every 1000 feet the beam travels, it spreads out by 1 foot. A mile away, the beam is 5 feet in diameter. The range also depends on receive antenna size and detector sensitivity — and, of course, upon the clarity of the atmosphere. I figured that the theoretical range of a 1-milliwatt, 1-milliradian laser and a photomultiplier receiver with a 3-inch lens would be 624 miles. In real life, atmospheric turbulence would limit this. Nevertheless, it's fair to say that if you use a decent receiver, the range is quite far.

Aiming a laser on a target several miles away is no small problem. Turning a 1-milliradian divergent laser by only 1 degree will move the 5-foot diameter beam almost 100 feet at a receiving site 1 mile away.

The laser must be mounted on a very stable tripod. I fashioned a micrometer-type mount to provide fine aiming adjustments. A rifle scope mounted on the laser will also help.

I made several contacts with the K6MEP contesters when I was with the Los Padres Microwave group. I was usually on Mt. Pinos, while MEP was on Reyes Peak, about 15 miles away. I used a 4-mW homebrew laser and K6MEP used a Spectra Physics laser owned by W6OAL. Both my unit and the MEP device used chopper modulation. I built both receivers around 931A photomultipliers. One used a 10-inch Fresnel lens (**Photo G**) and the other a 3-inch glass lens (**Photo H**). We made contact by slowly sweeping the lasers back and forth across what we believed to be the receiving site (the target was too far away to recognize by eye). The receiving site gave a yell on 2 meters when they saw the beam flash by. (The beam at that distance at night appears very bright; it is even visible in daylight.) Then the fine tuning began.

a word of caution

Do be very careful when using lasers. They can cause blindness. Don't ever point one at aircraft; it's not only dangerous, but illegal. Several people have been arrested for doing just that.

Special glasses and goggles that block the passage of the red output of Helium Neon lasers area available and are strongly recommended for anyone who experiments or works with lasers.

Compared to the other Ham bands, the usefulness of the world above 300 GHz certainly has its limits. But the purpose of Amateur radio is not to find the most efficient band, the most practical mode, and the most expedient QSO form. We are here to explore, to experiment, and to investigate.

As it is with the 432-MHz and higher bands, most lightwave communications activity will probably center on ARRL VHF/UHF and microwave QSO parties

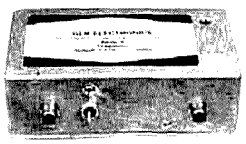
and contests. But undoubtedly some practical uses will be found. One might be for links between split-site repeaters, for example.

The age of the inexpensive laser is here now. Receivers are simple to build. Effective and affordable modulation methods are all that remains to be developed. Perhaps one of the readers of this text will discover a viable technique.

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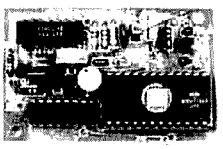
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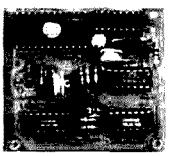
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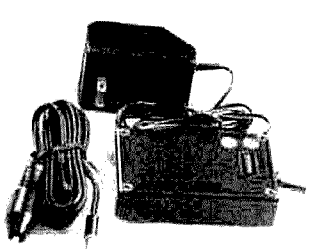
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advanced CW processor

Enhance reception with limiter/filter building blocks

The proper sequence and configuration of voltage amplitude limiters, bandpass filters, and catalytic carriers can enhance CW reception. **Figure 1** shows a series of blocks I have cascaded that work well together. At first glance this sequence may seem strange, but let me explain. To understand the reasons that these blocks help each other, it's best to start from the output end and work toward the input — excluding the power amplifier, which is conventional.

carrier-activated limiter

A technique referred to in part of **fig. 1** as CAL, for Carrier Activated Limiter, was completely described in a previous article.[1] The heart of this device is a wideband limiter that's constantly switched into upper and lower saturation by a high-frequency square wave carrier. Output from the limiter is then fed to a filter that passes a desired signal frequency and removes the square wave carrier. The carrier serves only as a catalyst to create a system that provides an S-shaped nonlinear gain characteristic with a well-defined output amplitude level. **Figure 2** shows the basic performance curve.

With this system, signal and/or noise voltage amplitudes only a few dB below the nominal 0.5-volt p-p designed signal level requirement will begin to be suppressed. If signal and/or noise voltages are 10 dB or more below the 0.5-volt level, they will be suppressed a nominal 14 dB below what they would otherwise be. In a similar manner, the ringing residues caused by the 100-Hz filter, when confronted by the typical rise time of a keyed CW signal, will also be suppressed from a normal 15 to 20 dB below the "dit—dah" level to around 30 or 35 dB below. Because of this feature, it is actually possible to reduce the preCAL filter bandwidth to as little as 20 or 30 Hz without losing the ability to copy relatively high speed CW: you get the essential capability of synchronous detection without its complexity.

The referenced article also carefully explains why the narrow filter in front of CAL is necessary to reduce the problem encountered when more than one signal enters any amplitude limiter function. But also consider what happens when the 100-Hz bandwidth filter is struck by a noise pulse.

pulsed 100-Hz bandwidth filter

This filter is included in the CAL article. **Figure 3** shows the output that results when it receives a noise pulse from the 500-Hz bandwidth filter. You'll note that the peak ringing amplitude from the 100-Hz bandwidth filter is about 1/6th that of the short pulse that caused it. This phenomenon is present for impulse types of noise any time you have a narrow filter following a significantly wider bandwidth system. What basically happens is that a tall, short-duration burst of energy is transformed by the narrower filter to become a low-amplitude, long-duration burst of energy that rings mostly as a sinusoid at the narrow filter's natural geometric center frequency. However, because of our logarithmic hearing characteristic, we cannot detect much difference between the two — perhaps, in part, because the total energy content difference between the input and resulting output pulses isn't nearly as great as the difference in voltage amplitudes.

But the CAL system doesn't care about energy content. It cares only about instantaneous voltage amplitude; so, for example, if a 0.5-volt p-p pulse arrives at the narrow filter's input, its resulting output of less than 0.1-volt p-p is low enough to be reduced by more than 30 dB by the CAL system. This illustrates why the present system already does quite well in controlling noise pulses that are no more than one or two times a properly adjusted signal amplitude. However, when confronted with noise pulses whose amplitudes are above three or four times the 0.5-volt p-p level, little suppression results. The stretched noise pulse is held to the maximum output level of the CAL design,

By Don E. Hildreth, W6NRW, 936 Azalea Drive, Sunnyvale, California 94086

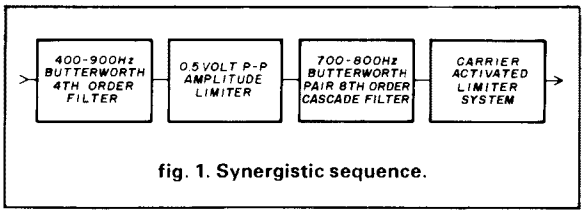


fig. 1. Synergistic sequence.

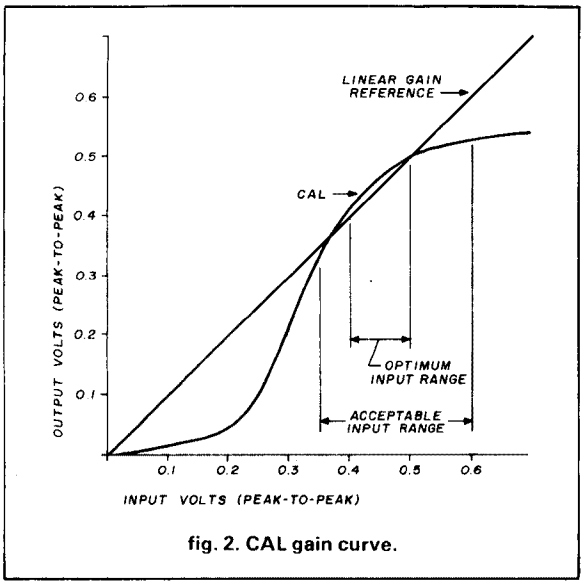


fig. 2. CAL gain curve.

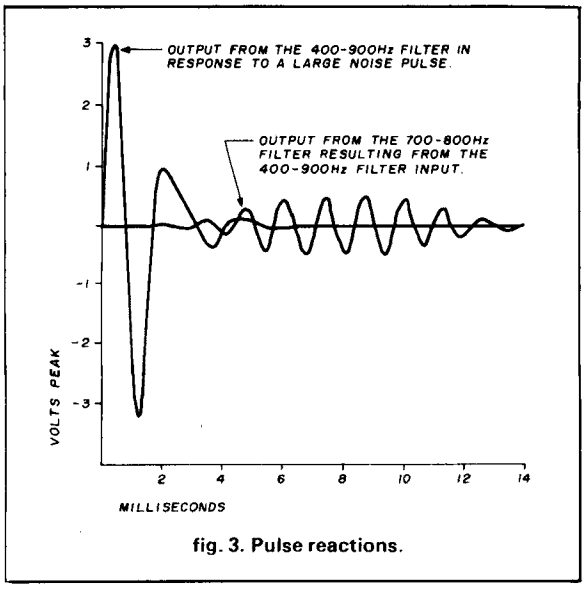


fig. 3. Pulse reactions.

but it can still be heard. Now, suppose a voltage amplitude limiter set at a limit level of 0.5 volts p-p is placed in front of the 100-Hz filter.

limiter addition

The CAL article carefully documents what happens if more than one signal enters a limiter. But how about those times when QRM isn't too heavy? During these opportune times, a front limiter can force noise pulses of very large amplitudes to be held to a value that can be acted on by the CAL's suppression factor. At the same time, a desired signal at 0.5 volts p-p slides right under the "limiter, filter, CAL" system unscathed. However, if a full 3 kHz to 5 kHz or so of audio bandwidth from your receiver is connected to this first limiter, then the number of opportunities for using the limiter may not be too favorable. Now we get to the 500-Hz bandwidth filter.

adding an offset first filter

Now this may seem a bit strange. A 400- to 900-Hz filter will pass a signal to the 750-Hz, 100-Hz bandwidth filter because it embraces that band, but its center frequency is 600 Hz, the geometric center of the band edges. Its residual ringing will be at around 600 Hz, which is outside of the 700- to 800-Hz passband of the 100-Hz bandwidth filter. The combined effect of the wider bandwidth and the offset center frequency enables a filter — restricted to only a 500-Hz bandwidth — to provide the 1/6th impulse-to-ringing signal amplitude ratio out of the 100-Hz bandwidth filter, as is shown in **fig. 3.**

An empirical equation used to design this system is:

$$BW1 = 4(1 + n/8)\ (BW2) \qquad (1)$$

Where BW1 is the 3-dB bandwidth of the first filter and BW2 is the 3-dB bandwidth of the second filter (narrow 8th-order filter).

This calculated minimum usable bandwidth for the first filter is used simply to increase the potential percentage of time when the first limiter can be used. The first limiter can be used with good results directly between your receiver and the 100-Hz bandwidth filter for impulse noise control, but when signal density is high it will often have to be switched out, especially if you are listening to a weak signal that requires high gain from your receiver. The filter shown here provides what I believe to be the best available compromise between potential use time and the degree of impulse noise suppression when the CAL is included. In addition to the use factor, another trade-off must be considered because of the more likely noise sources.

noise types

The efficacy of a noise limiter is strongly influenced by the preceding filter's bandwidth because of the length of time by which a filter stretches impulse noise. An empirical approximation of this for the main energy burst coming from a filter following an exciting pulse is given by the equation:

$$\text{Output pulse duration, } t_p = \frac{(1 + 0.5n/8)}{BW} \text{ seconds}$$

(For flat top filters with six, eight, or more poles, the undulation ringing can greatly exceed t_p given by this equation; see reference 2).

This filter behavior tends to place a limit on how close together in time noise pulses can be before the system starts to become ineffective. If, for example, a train of pulses that are too close together to allow the filter response to decay between pulses comes along, then a continuous output will build up at the filter's center frequency (this implies another benefit brought about by offsetting the first filter's center frequency). In this design, the first filter's t_p is just slightly longer than 2 milliseconds, so it cannot easily be undone by two of our most ubiquitous noise sources: the woodpecker and those offensive light dimmer controls. The repetition rate of the woodpecker is fairly slow and, although its pulse duration is longer than most noise forms called "impulse," it is still short enough to be suppressed by this system. Dimmer control pulses are very short when they output from a typical receiver, but they come twice per cycle at the 60-Hz power line rate, or every 8.3 milliseconds. To avoid this prevalent noise type, the first filter should have a bandwidth preferably three or four times 120 Hz and should avoid a center frequency at or close to three times 120Hz in particular. Calculations indicate that the system is effective against most automobile ignition noise up to 2000 or 3000 RPM for most modern autos. If you're in an area with heavy, multiple-auto ignition noise, it may be advisable to increase the bandwidth of the first filter in this system.

Noise from those devices that pass current through brushes to an armature, such as some motors and generators, probably can't be helped because the arcing often produces very high repetition rate random impulse-like electrical noise.

Figure 4 is a detailed schematic of the complete sequence.

operating strategies

I have always tried to limit the number of controls in any add-on device as much as possible. But if the number of controls must grow, at least most of them should be functions that require infrequent attention. In the CAL system, marketed as Model 10, there is just a *Mode* control and an output power level control, which are rarely adjusted. The system shown here adds only a toggle switch. For me, it has resulted in a significant improvement: I can now have ear protection in any position of the mode switch. If I'm listening to a strong signal that isn't being battered about by QRM, I use the wide band position, include the first limiter, and adjust the input level from my receiver to

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place the signal just at the edge of limiting. If QRM appears, I switch to the 100-Hz bandwidth position, then if — and only if — my desired signal is being "dimpled" by the interference, I toggle out the front limiter. I bring in the CAL only if a signal is quite weak, or if the keying is fast and I want to reduce ringing.

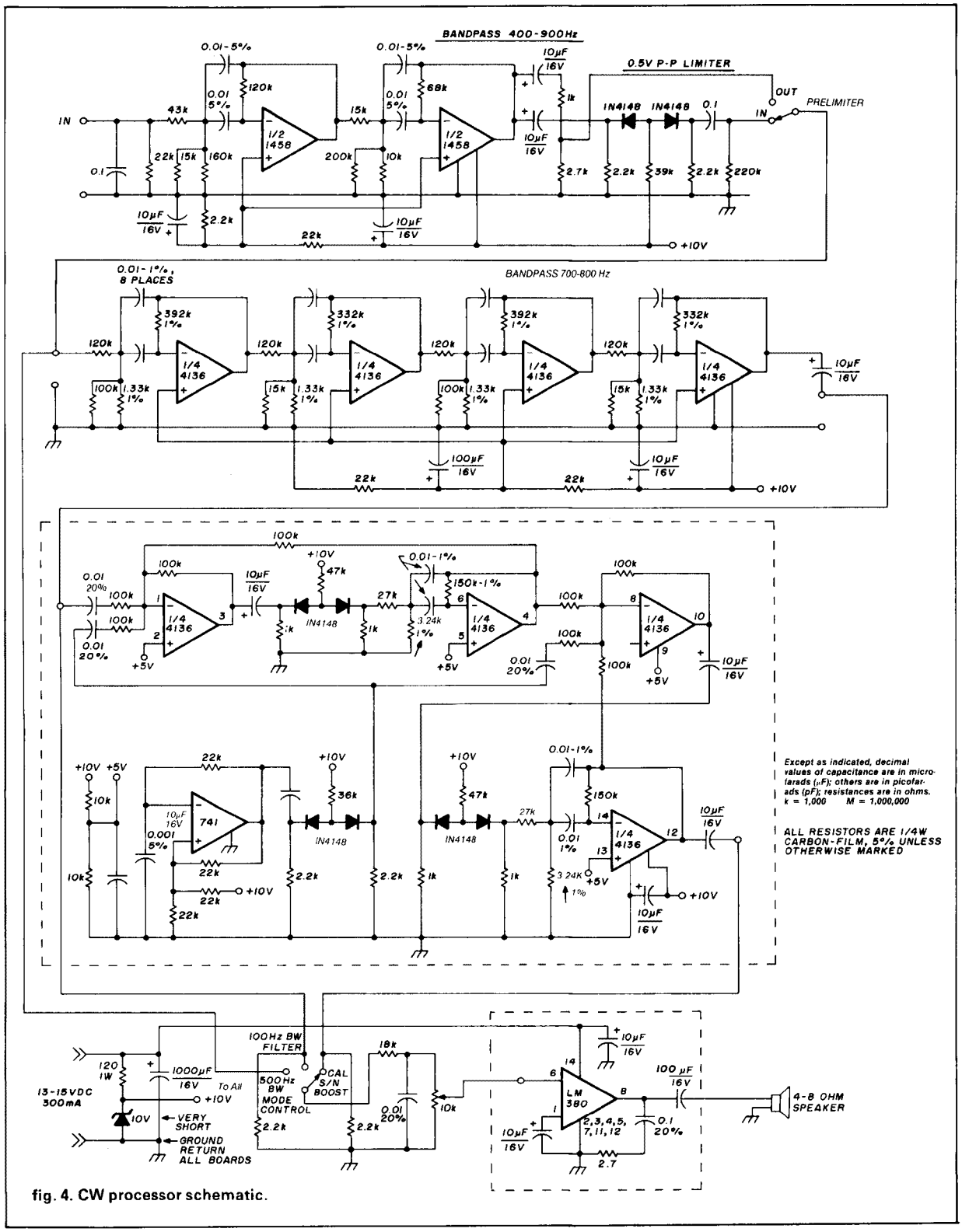


fig. 4. CW processor schematic.

Receiver AGC is off and/or defeated by a moderate-to-low rf/i-f gain setting.

As you can imagine, using the CAL for a weak signal with QSB can be quite a chore — and sometimes impossible (see **fig. 2**). If not too much QRM is present in the 500-Hz bandwidth filter, the first limiter can iron out this QSB very nicely by advancing the signal level from your receiver to absorb the variations. By this action, the signal level fed to the CAL is forced to remain in the optimum region. Using these techniques, I've been able to copy some DX signals that I couldn't read any other way.

final comments

So what about the bad news? Certainly there always has to be some bad news. Well, this is it: because the CAL system suppresses signals on the skirts of the 100-Hz bandwidth filter, the system starts to tune more like 70 or 80 Hz with extremely steep skirts. This can make tuning difficult unless you have plenty of bandspread, very little backlash, and a frequency-stable receiver. Using a Kenwood TS-820S, whose tuning rate is around 20 kHz per 360 degrees of knob rotation, care is required — but it's not difficult with a 100-Hz bandwidth filter. I found that the approximate limit for operation with this tuning rate was a 40-Hz bandwidth Gaussian filter in place of the 100-Hz Butterworth cascade unit. . . and knob control is very demanding with that!

Based upon my observations, the limit on how narrow our useful bandwidth can be for copying most CW signals is determined by the prevalent mechanical/electrical tuning (bandspread), frequency stability, and inherent bandwidth requirements based upon transmitted information rates. Limitations for most equipment are in the order given, with the mechanical problem about ten times worse than the stability factor and the stability limit about ten times worse than the information bandwidth limit. To test this, a special multi-pole 10-Hz bandwidth filter was prepared. Copy was nearly impossible with the straight filter, but quite good when using CAL. Tuning was next to impossible, but once a signal was tuned, it usually remained in the window for the duration of a typical QSO.

Now that frequency synthesizers have found their way into our gear, the 10-Hz bandwidth will become more and more practical. To make these narrow filters easier to work with, and for more other subtle reasons, visual aids to aural-only detection should be applied.[3] One thing I can tell you for sure: the bands seldom seem crowded when you're using these very narrow bandwidths!

references

1. Don Hildreth, W6NRW, "Carrier-Activated CW Reception Limiter," *ham radio*, September, 1985, page 113.
2. Don Hildreth, W6NRW, "Graphic Filter Design," *ham radio*, April, 1984, page 37.
3. Unpublished manuscript; please send SASE for copy.

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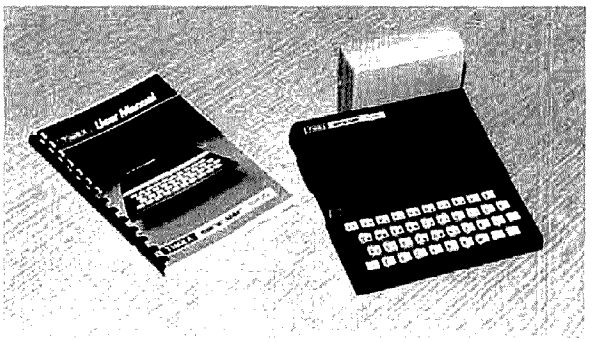
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the ZX-81/TS-1000 controller:
new use for old computers

Four ICs in interface circuit link computer to outside world

If you've been to a hamfest lately, or read the back pages of any ham magazine, you've probably seen used Sinclair ZX-81 or Timex/Sinclair 1000 computers selling for less than $20*. During their heyday several years ago, these computers — which originally sold for $150 — led the way for the personal computer revolution. Although the Sinclair was somewhat of a marvel at the time, many users soon became disenchanted with it, either because of the small membrane keyboard or the difficulty of interfacing it with peripherals.

Perhaps you've stashed a Sinclair away in your junkbox in hope of someday finding something to do with it. Or perhaps you've toyed with the idea of picking one up at a hamfest but decided otherwise. Never mind. *Get one.*

This small, cheap-looking computer has computing power comparable to much more expensive computer systems. The key to harnessing this power is simply finding an application that takes advantage of its strengths. For while it's admittedly disappointing for some applications — such as word processing — it's ideally suited for control functions and can be easily programmed for a host of controller applications.

In this article, a simple circuit using only four ICs will be presented, along with a detailed example of how to program your ZX-81 or TS-1000 computer for I/O applications.

With this simple interface circuit attached to the memory port of either the ZX-81 or TS-1000, you instantly gain the ability to communicate with the outside world. All sorts of applications — including satellite rotor controls, digital voltmeters, morse code sending and receiving, educational demonstrations, and many others — easily follow once you have an I/O interface. I developed this circuit as part of a school project a few years ago when I used the TS-1000 computer to send and receive RTTY. Since then, I've been selling the information contained herein at cost to Sinclair computer enthusiasts at local hamfests and by mail. If you've been wondering how to make your ZX-81 or TS-1000 perform some useful tasks and are interested in a neat one-evening construction project, then read on.

what is a controller?

While many computers are used for computation, word processing, or data management, many "small" computers, composed of just a microprocessor and a handful of special-purpose chips are built strictly for *controller* applications. For example, microprocessor circuitry is now being installed in automobiles. A microprocessor is programmed with a simple looping program which does nothing more than continuously monitor various data lines and then branch to a par-

By Ted S. Rappaport, N9NB, Box 283, Electrical Engineering, Purdue University, West Lafayette, Indiana 47907

*TS 1000 computers ($19.95), manuals ($2.95), and accessories are available from Hal Tronix, Inc., Dep't HR, P O Box 1101, Southgate, Michigan 48195 (313 285 1782).

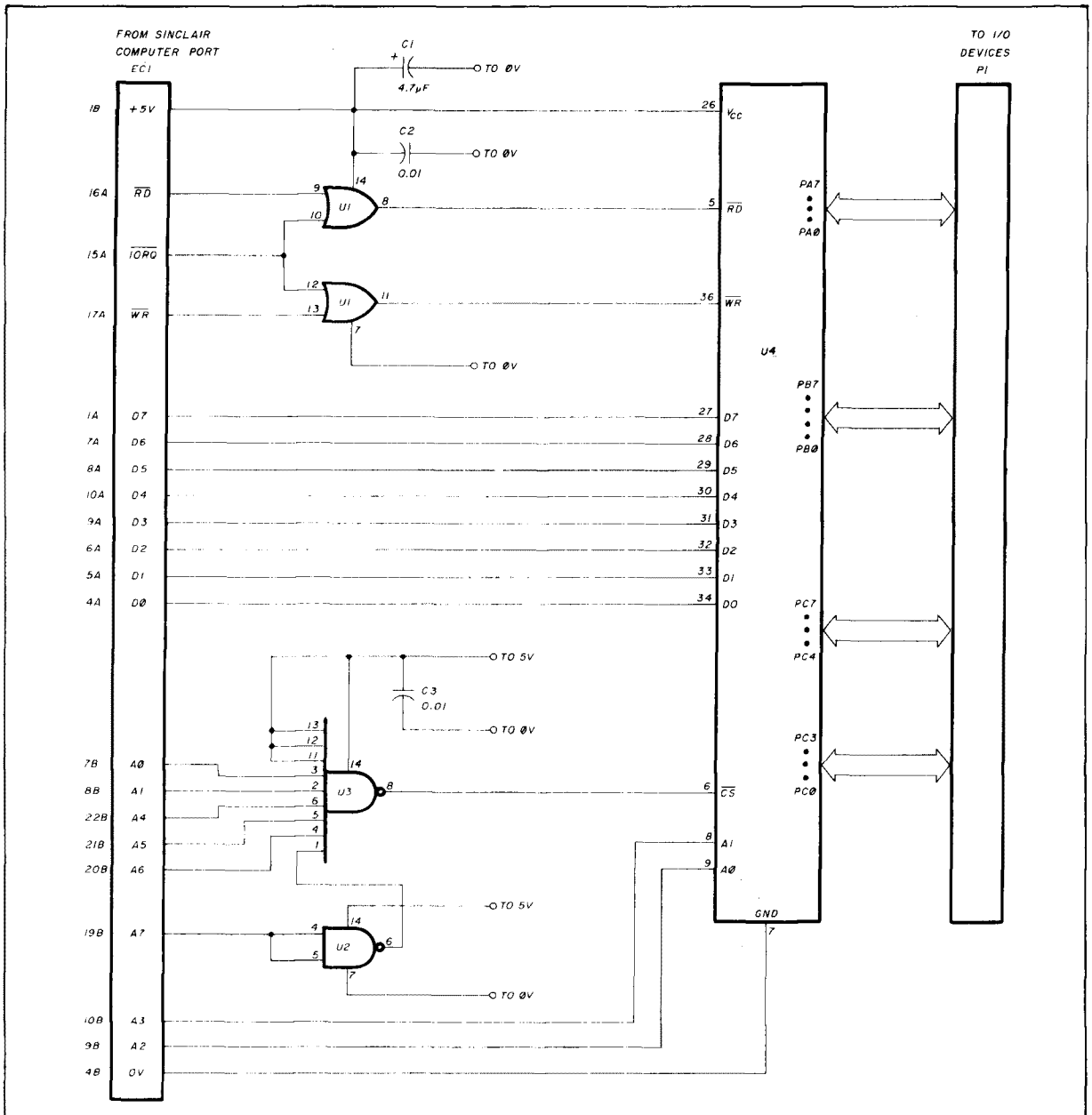


fig. 1. Schematic diagram for interface circuit.

ticular part of the program when specific lines change state. These input/output (I/O) lines are connected to hardware such as the fuel level sensor, the climate control console, and thermal sensors under the hood and inside the passenger compartment. When the driver wants to raise the car's interior temperature to 75 degrees, he or she simply presses a button, entering the desired temperature. The program in the microprocessor is continuously looping, searching for any changes on the input data lines. Finding that the driver has activated the climate control console line, the program then branches to the subroutine designed to effect the desired climate change. In this case, the information (75 degrees) is read into the microprocessor as digital data and a control signal is then sent out to the appropriate hardware (in this case, the heater on/off switch). While this is an overly simplified version of what is actually done in today's automobiles, it illustrates one typical application of a controller.

Of key importance is the fact that in such applications, the computer is programmed to handle a variety of input signals from external hardware and issue

Table 1. Parts list for Interface Circuit (fig. 1).

Schematic Designator	Description
C1	4.7 μF electrolytic capacitor
C2	0.01 μF capacitor
C3	0.01 μF capacitor
EC1	2 row x 23-pin edge connector; 0.1-inch spacing, pin length at least 0.75-inch.
P1	DB-25 D-connector
U1	74LS32 Quad two-input OR chip
U2	74LS00 Quad two-input NAND chip
U3	74LS30 Eight-input NAND chip
U4	8255 Programmable Interface Chip

Miscellaneous:
Single-sided circuit board
Jumper wires (at least 10)
Project box (UNIBOX UB270-130)

control signals to external hardware. Because the controller's program is simple and repetitive, it doesn't require much memory. The 2K of memory in the ZX-81 and TS-1000 is more than enough to program even a very complex controller. We can also take advantage of the computer's high-level (BASIC) capability to simplify programming the computer for controller applications. An understanding of machine language is important, though, since the I/O interface must allow exchange of data on a level lower than BASIC; this is done through the *PEEK* and *POKE* commands.

description of interface circuit

With the interface circuit connected to the port of the ZX-81 or TS-1000, very fast data signals which appear on the data bus of the computer can be latched by the interface circuit and sent as control signals to external hardware (I/O devices). Similarly, in conjunction with the interface circuit, the computer can read in very fast asynchronous signals from external hardware and use the received signals to make decisions in software.

Figure 1 is the schematic diagram of the I/O interface. A complete parts list appears in **table 1**. Note that no power supply is needed for the interface; all four ICs are conveniently powered by the computer's internal 5-volt supply (the voltage is made available at the computer port).

The heart of the interface circuit is U4, the 8255 programmable peripheral interface (PPI) chip. U1, U2, and U3 serve to decode the address signals generated by the computer. After much trial and error, I found that it's possible to access four I/O address locations not used by any of the internal computer circuitry. These address locations — 115, 119, 123, and 127 (decimal) — may be used to activate any external device connected to the computer port. For my interface circuit, these four address locations are used to access U4, which in turn directs the flow of data between any external I/O device and the computer. In **table 2**, I've shown the addresses used to control the various registers of U4. While some of the basics on how to use U4 will be spelled out later, more details can be found in any Intel databook or in Goldsbrough's *Microcomputer Interfacing with the 8255 PPI Chip*.[2]

Table 2. Operation of 8255 PPI chip (U4).

I/O Instruction	I/O Address	Operation
IN	115	Read from Port A
IN	119	Read from Port B
IN	123	Read from Port C
OUT	115	Write to Port A
OUT	119	Write to Port B
OUT	123	Write to Port C
OUT	127	Write to Control Register

When laying out a circuit board, I decided that it would be nice to have the ability to add components or connectors in the future. (I call the upper portion of the board, where there are several inches of 0.1-inch spaced holes, the "playground" area.) By adding components and appropriate jumper wires, it's possible to expand the interface circuit. The single-sided board negative I used, shown in **fig. 2**, requires ten insulated jumper wires, which should be installed on the foil (non-component) side of the circuit board after all of the components are mounted (see **table 3**). Pads are available on the circuit board to permit easy soldering of each jumper. I used 16 additional jumpers to route all eight of the port A pins, four of the port B pins, and four of the port C pins from U4 to molex pins attached on the top of the "playground" area. These molex pins mate with headers that allow for easy detachment of the data lines from I/O hardware or from a case-mounted connector. You may wish to use just one pin or all 24 pins available on U4 for I/O data exchange. Use as many data lines as you need.

The component layout for the circuit board (**fig. 2**) is shown in **fig. 3**. When mounting the ICs, be sure to use a low-wattage soldering iron since the pads are delicate. I strongly recommend the use of sockets for each of the four ICs.

All of the components are easy to acquire at hamfests or electronic shops, except, perhaps, for the 46-pin connector used to attach the interface board to the computer memory port. The connector type is the same as that used on ZX-81 and TS-1000 peripherals such as the 16K RAM-pack, and is a 46-pin, double-sided edge connector with 0.1-inch spacing between contacts. I found that you have to buy a sturdy con-

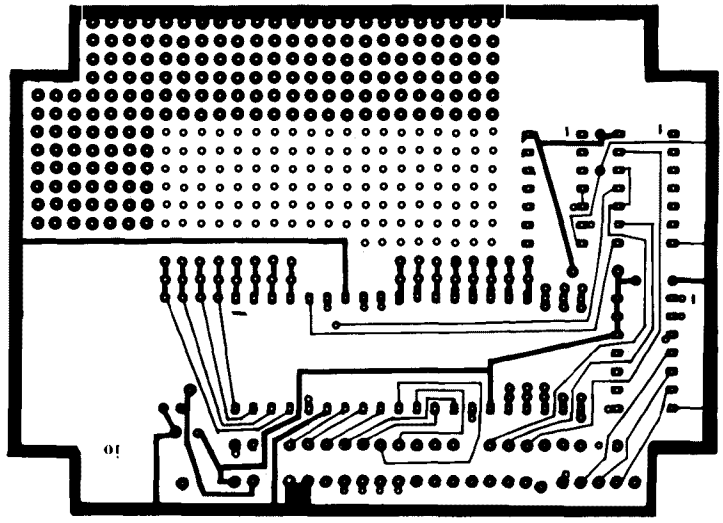


fig. 2. Single-sided circuit board artwork for I/O interface circuit.

Table 3. Jumper wires needed for single-sided circuit board. These connections should be made after all other components are soldered in place.

Jumper between: U1, pin 11 and U4, pin 36
Jumper between: U2, pin 5 and EC1, pin 19B
Jumper between: U2, pin 6 and U3, pin 1
Jumper between: U3, pin 2 and EC1, pin 7B
Jumper between: U3, pin 3 and EC1, pin 8B
Jumper between: U3, pin 8 and U4, pin 6
Jumper between: U3, pin 16 and U1, pin 16
Jumper between: U4, pin 8 and EC1, pin 10B
Jumper between: U4, pin 9 and EC1, pin 9B
Jumper between: U4, pin 27 and EC1, pin 1A

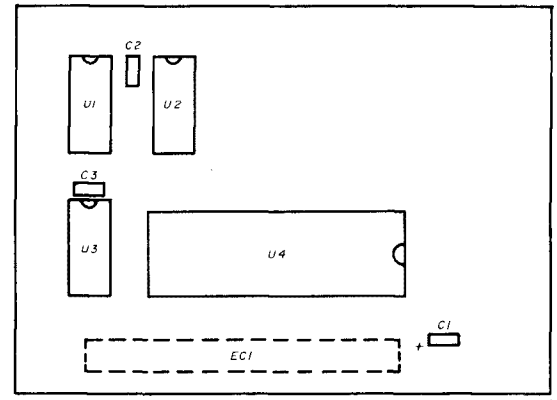


fig. 3. Component layout for interface circuit — view from component side (EC1 mounted on FOIL SIDE).

nector and be prepared to use a Dremel tool or a small saw to cut a larger connector down to correct size. But be careful: I bought a 60-pin, double-sided connector for less than $1.00 from a military surplus dealer at a hamfest. Eagerly, I took the new connector home, cut it to length, and soldered it onto the board. After a couple of trial fits on the memory port of the computer, the connector crumbled into tiny bits! It was a real pain to have to chop off the faulty connector and resolder a new one onto the board! You may notice from the accompanying photographs that the connector must be soldered on the foil side of the board, and protrudes from that side of the board as well.

The entire cost of components is well under $20. The construction time is fast — one evening should be sufficient.

I was able to find attractive plastic boxes made by UNIBOX[3] at a local Radio Shack for less than $4.00 each. The circuit board fits neatly inside, and only two rectangular holes have to be cut into the box: a 0.5 x 2.5-inch hole to allow the 46-pin edge connector to protrude from the circuit board, and a 0.5 x 1.0-inch hole on the other side of the box to allow I/O wires to be fed to a D-connector mounted on the exterior (see **fig. 4**).

interface software

You can add all the hardware in the world to a computer, but it won't serve any useful purpose unless you know how to program the computer effectively. In this section a programming example for an application of the interface circuit is given, and a review of how to

use the ZX-81 and TS-1000 *PEEK* and *POKE* commands is presented.

In order to perform data exchanges between I/O devices and the computer, it's first necessary to set up the I/O format on U4. U4 has 24 pins which are usable for data exchange. These pins are partitioned into three groups of eight pins. Each group of eight pins is called a *port.*

Any program written for I/O control should begin with an initialization stage. This initialization sets the data flow of the ports on U4. For example, you may want to program the eight data lines of port A for output (data sent to the I/O device from the computer) and have the eight data lines of port B programmed for input (data sent from the I/O device to the computer). While ports A and B require that all eight pins be either input or output, Port C allows for four pins to be input and four to be output. Of course, Port C may also be programmed for all eight pins to be either input or output.

While it's beyond the scope of this article to teach assembly language and machine code programming, let me assure you that mastering them is not difficult. If you find the example in this article insufficient, you may wish to read Chapters 1, 2, and 5 in Hordeski's easy-to-read *Microprocessor Cookbook.*[5]

The following example details the simplicity of controller software. Consider an alarm system that monitors the status of eight windows around the house. Three wires (ground, +5 volts, and data status line) could be routed to each of eight windows. A small switch could be mounted by each window so that if the window is shut, a low-logic level (0 volts) exists on the status line. If any of the windows are open, we want the computer to recognize this and sound a bell. For this case, we may use all eight pins of Port A for input (to read the status of each of the eight windows), and may use Port B for output (although we need to use only one of the eight pins to output a control signal for the alarm). Since we're not using Port C, we don't care what it's programmed for! By connecting the status line of each window to individual pins of Port A, we can read the status of each window into the computer. If we use a simple transistor switch to activate the bell, one of the pins from Port B can be connected to the gate of the transistor switch. When a window is discovered to be open, the computer then sends out a high-logic level (5 volts) to the alarm pin (as well as the unused pins) on Port B, thus sounding the bell.

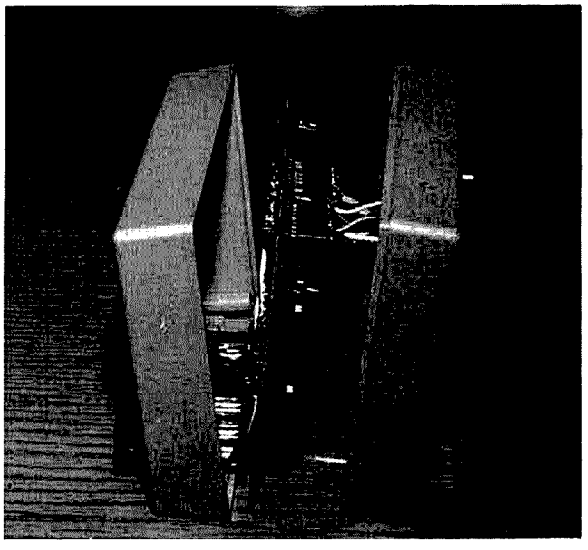


fig. 4. The circuit board fits into the project box. Be sure to use an edge connector with at least ¾ "of pin length to allow clearance when the circuit is plugged into the computer port. It is convenient to have a connector on the back of the box for easy access to the I/O lines.

The programming steps for the window alarm example would be as follows:

1. Initialize data flow on Ports A, B, and C.
2. Look at input data from windows.
3. If any of the input data bits is "1," sound alarm.
4. If the input data is all "0," then go to step 2.

Notice that once the data flow is initialized on U4, no further action is required unless the power is unplugged or U4 is reprogrammed. The assembly language program for this example follows (comments are included in parenthesis):

```
(Initialization):
LD A, 144 (144 = 10010000. This programs Port A for
input, Ports B, C for output. We load 144 into ac-
cumulator.)
OUT 127,A (Send contents of accumulator 144 to the
U4 control register, which is located at ADDR 127.)

(Monitor window data):
loop  IN A, 115. (Read in the eight-bit data at ADDR
      115 — Port A — and place data in the ac-
      cumulator.)

JRZ, loop  (Compare the eight bits of data in accumu-
           lator to the all-zeros data word 00000000;
           if all eight bits of data are zeros, then jump
           back to loop and continue to monitor.)

(If program gets to this point, then sound the alarm!):

LD A, 255 (Load accumulator with all-ones word,
          11111111)

OUT 119, A (Send the accumulator contents out to
           Port B; this sounds the alarm).
```

The preceding assembly language program, when written in machine code, requires only 13 bytes of memory! It's entered into the computer via the *POKE* command. The conversion from assembly language mnemonics to machine opcodes is given in the ZX-81 and TS-1000 user's manual appendix. For the previous program, the machine code that would be POKEd is given in **table 4**.

Although not illustrated in this example, it's simple to monitor your I/O activity by going back and forth between BASIC and machine code. In this case, it would have been easy to see on the computer monitor which particular window (or windows) were opened by simply passing the Port A data to the b-c register pair in the machine code program, and then by returning to BASIC by way of the USR command and printing the value of the USR function on the screen.

Table 4. Machine code for programming example.

Assembly language instruction	Machine code
LD A, 144	62
	144
OUT 127, A	211
	127
IN A, 115	219
	115
JRZ, -4	40
	252
LD A, 255	62
	255
OUT 119, A	211
	119
RETURN TO BASIC	201

machine code within BASIC programs

To store a machine code program in the computer, you must first set aside some memory by using a REM statement (usually the first line in your BASIC program). The REM statement sets aside successive memory locations which are never written over, thereby allowing you to store machine code programs safely. Machine code programming may be saved on tape by using the *SAVE* command.

A machine code program is run in sequence by using the USR command, where the argument of the USR command is the address location of the first machine language opcode. The computer returns to the BASIC program once the return opcode (201) is encountered, and assigns the decimal value of the b-c register contents to the variable name to which the USR function is equated. In this manner, it's possible to pass data from machine code programs (and consequently data from external I/O devices) to BASIC. Or you may use the *PEEK* command in a BASIC program to retrieve data from dedicated memory locations in your machine code storage area (REM statement). Either way, it's a simple chore to transfer data from "real world" devices to your computer. By using the *POKE* command in a BASIC program, you may place data from BASIC into data locations of machine code programs. Note that this allows you to make keyboard entries and have these entries correspond to control signals for I/O devices.

When you're ready to enter a machine code program into the computer, clear the computer's memory and begin as if you were writing a new BASIC program. Enter **line 1** as a REM statement, filling in as many letters in the REM statement as the number of bytes required in your machine program. It doesn't matter what you put in your REM statement, because after you *POKE* in your machine code, the REM statement appears as gibberish on the screen. For the TS-1000 computer, the first usable memory slot once a REM statement is entered is 16514. Use the *POKE* command to place the machine code in sequential memory locations starting with 16514. Upon completing the entry of your machine code, place the opcode 201 in the very next memory location following your last program opcode. At this point it's a good idea to save your program on tape (even though it's only a single REM statement). That way if anything happens — for example, if junior unplugs your computer — you don't have to re-*POKE* your entire program! For the machine code in **table 4**, the following would be entered into the computer:

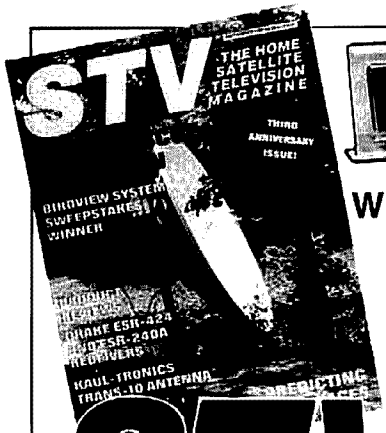
```
POKE 16514,62 (press enter)
POKE 16515,144 (press enter)
POKE 16516,211 (press enter)
. . . .
. . . .
. . . .
POKE 16524,211 (press enter)
POKE 16525,119 (press enter)
POKE 16526,201 (press enter)
```

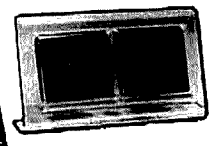
To verify that the opcodes are in memory, you may use the *PEEK* command to see what the computer has stored.

In order to execute your machine code, you must have a BASIC statement that contains the USR function. For the window alarm program, your next program line (following the REM statement) could be:

2 LET WINDALARM = USR (16514)

By running this simple two-line BASIC program, your Sinclair computer executes the machine code and is able to communicate with external devices through the use of the interface circuit! If the machine code was written so that the Port A data was placed in the b-c register pair, then the value of the variable WINDALARM, which can be viewed on the monitor with a simple print statement such as





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could be displayed to indicate (in decimal) the window status.

When using the Sinclair computer as a controller, more than one machine language program can be used. It's often convenient to write several *subroutines* in machine code, and store them all in succession in the first REM statement. (Remember they're separate to the computer when they end in the opcode 201). In this manner, your BASIC program can consist of various calls to the subroutines by using USR functions with the appropriate memory addresses in the REM storage.

conclusion

In this article I've discussed the hardware and software needed to provide a communication path between your Sinclair ZX-81 or Timex/Sinclair 1000 computer and any I/O device you might wish to control or monitor. With the price of these computers so low, even those with the tightest budgets can afford to experiment with computer control! By now, you may already have an idea for a controller application in your shack. If you take an evening out to build this circuit and spend a few hours reviewing machine language programming, you'll be amazed at what your small but mighty computer can do!

references

1. *Intel* is a trademark of Intel Corporation. Information on Intel products may be obtained from Literature Department, Intel Corporation, 3065 Bowers Avenue, Santa Clara, California 95051.
2. P.F. Goldsbrough, *Microcomputer Interfacing with The 8255 PPI Chip*, Howard W. Sams, Publisher, 1979.
3. UNIBOX, Amerex Corporation, Riverside, California 92502.
4. Michael Hordeski, *Microprocessor Cookbook*, Tab Books, 1979.

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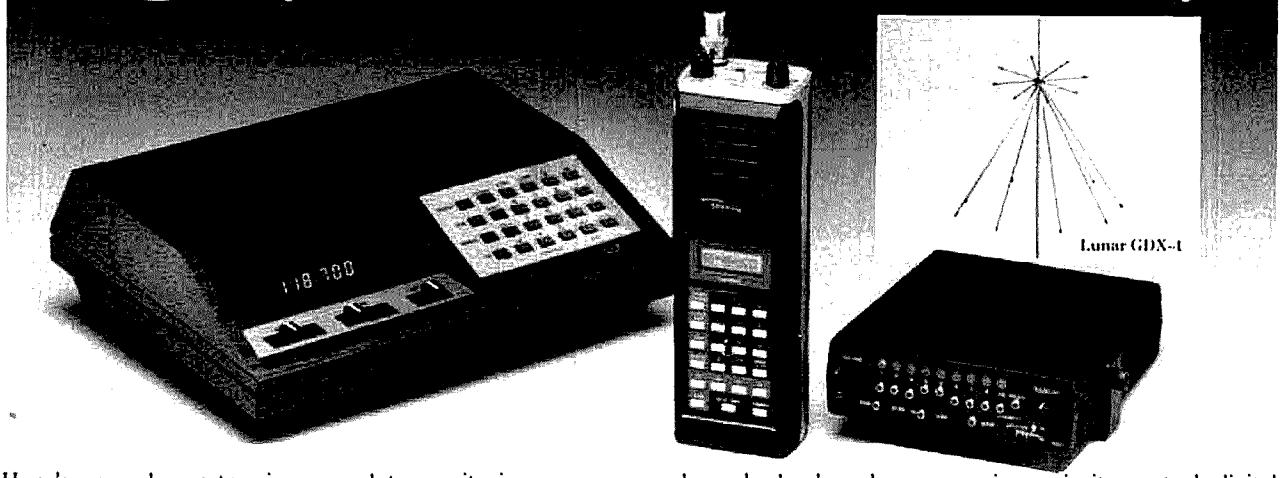
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amateur FSK: a spectral analysis

Experimenting with spread spectrum? Try this technique to assess interference levels and determine channel separation

By Richard Ferranti, WA6NCX, 116 Franklin Street, Arlington, Massachusetts 02174

The increasing use of digital communications techniques in Amateur Radio, be it in RTTY, packet radio, AMTOR, or spread spectrum, makes it increasingly important for the Amateur to understand the characteristics of digital transmissions. This article discusses frequency-shift-keyed (FSK) waveforms, the simple technique of associating one frequency with one level of a digital code and another frequency with the other level. In essence, the binary code switches a carrier between two frequencies, a method well-known in the 1920's, when radio teleprinting first became popular.

This article investigates the spectral characteristics of FSK signals — that is, the way an FSK signal occupies a segment of the radio frequency spectrum. Spectral analysis is invaluable for assessing interference levels, deciding on adjacent channel spacings, and more recently for spread spectrum experimentation. Although two computer-simulated and experimental studies of some FSK signals have appeared in the Amateur press, neither study was based on mathematical analyses of the waveforms, and consequently encouraged little further investigation among the Amateur community.[1,2] Here, the mathematical results will be provided, thus encouraging readers to experiment with FSK waveforms on their personal computers.

generating an FSK signal

There are two basic ways to generate an FSK signal. Referring to **fig. 1**, we see two oscillators running at frequencies $f1$ and $f2$, with a switch to choose one or the other as an output. The switch is controlled by the ones and zeros of a digital bit stream: $f1$ is sent for a zero, $f2$ for a one. The resultant frequency-shift-keying waveform has discontinuous phase at the transition times, meaning that when a bit changes from one to zero or zero to one, the phase of the $f1$ oscillator is completely *unrelated* to the $f2$ oscillator phase. Some Amateur modems generate FSK this way, usually from two separate switched audio oscillators fed into the microphone input of an SSB transmitter.

In **fig. 1B**, the FSK waveform is generated by feeding the digital bit stream to a *single* voltage-controlled oscillator. The oscillator is instantly pulled from $f1$ to $f2$ and back again, but because there's only one oscillator the phase is continuous at the bit transitions. The old trick of FSKing the VFO of a transmitter by tying an element of the oscillator tube to the keyboard also yields continuous-phase FSK (CPFSK). Modern radio modems generating audio FSK using an integrated circuit like the XR2206 can also generate CPFSK. The difference between discontinuous phase FSK (DPFSK) and CPFSK is illustrated in **fig. 2**, where the bit stream and the two frequencies it controls are plotted. Note the sharp-edged transitions in DPFSK when the carrier switches between $f1$ and $f2$; these are not present in the smoother CPFSK waveform. As shown later, the spectral responses of these two kinds of FSK are quite different.

overview of spectral analysis

An easy-to-understand time-domain representation of FSK signals, that is, how they would look on an oscilloscope, is shown in **fig. 2**. However, it's much more interesting to consider what these signals look like in the frequency domain. Questions such as how much bandwidth they occupy, why some seem to splatter more than others, and what kind is better for radioteleprinting than for spread-spectrum communications are answered. It's important to realize that we have a couple of parameters to play with here — the

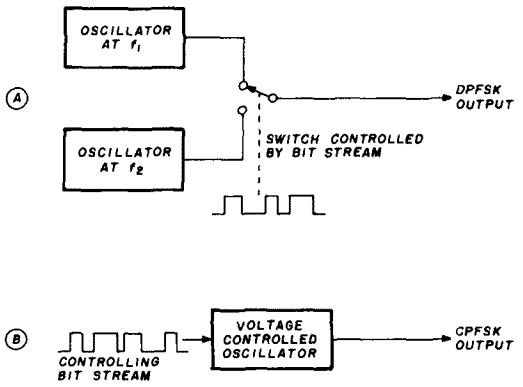


fig. 1. Two methods of generating frequency-shift-keyed signals.

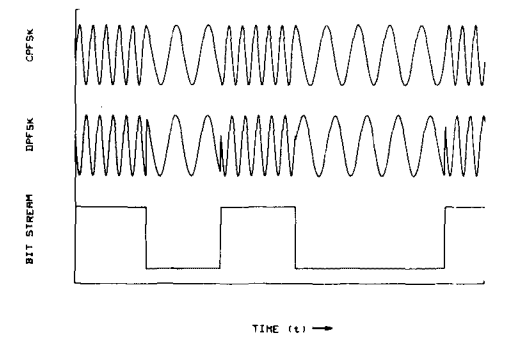


fig. 2. Continuous and discontinuous phase FSK as viewed on an oscilloscope.

```
5 REM **** PROGRAM FRAGMENT FOR
  CALCULATING SPECTRAL DENSITIES ****
10 DIM X(200),Y(200)
20 PI=3.14159265:T=0.022:SHIFT=170:W1=90
83.3*2*PI:W2=(9083.3+SHIFT)*2*PI:WS=(908
3.3+SHIFT/2-1000)*2*PI
30 AL=(W2+W1)/2:BE=(W2-W1)/2
40 FOR Z=1 TO 200
50 REM **** DISCONTINUOUS PHASE CASE
  --- USE LINES 60 TO 120 ****
60 A=WS+Z*20*PI-W1:B=WS+Z*20*PI+W1:C=WS+
Z*20*PI-W2:D=WS+Z*20*PI+W2
70 A1=(SIN(A*T/2))*(SIN(A*T/2))
80 B1=(SIN(B*T/2))*(SIN(B*T/2))
90 C1=(SIN(C*T/2))*(SIN(C*T/2))
100 D1=(SIN(D*T/2))*(SIN(D*T/2))
110 E=A1/(A*A)+B1/(B*B)+C1/(C*C)+D1/(D*D
)
120 Y(Z)=10*CLOG(E):X(Z)=(Z-100)*10
130 REM **** CONTINUOUS PHASE CASE
 ---- USE LINES 140 TO 280 ****
140 W=(2*PI*Z*10)+WS
150 A=W-W1
160 B=W-W2
170 C=W+W1
180 D=W+W2
190 E=W-AL
200 F=W+AL
210 G=2*SIN(A*T/2)*SIN(A*T/2)*SIN(B*T/2)
*SIN(B*T/2)
220 H=2*SIN(C*T/2)*SIN(C*T/2)*SIN(D*T/2)
*SIN(D*T/2)
230 I=(1/A-1/B)*(1/A-1/B)
240 J=(1/C-1/D)*(1/C-1/D)
250 K=T*(1-2*COS(E*T)*COS(BE*T)+COS(BE*T
)*COS(BE*T))
260 L=T*(1-2*COS(F*T)*COS(BE*T)+COS(BE*T
)*COS(BE*T))
270 M=G*I/K+H*J/L
280 Y(Z)=10*CLOG(M):X(Z)=(Z-100)*10
290 NEXT Z
```

fig. 3. Atari BASIC listing for calculating the spectral density of FSK signals. The trig functions evaluate their arguments in radians (not degrees).

words per minute or baud rate, and the frequency shift (from $f1$ to $f2$ in Hertz).* For instance, an 850-Hz shift, 60 wpm Baudot FSK signal has a different spectrum than the same signal shifted by only 170 Hz. By exploring different shifts and data rates, it's possible to optimize FSK for one's particular needs.

The time-domain representation of an FSK waveform is converted to a frequency spectrum by a mathematical operation known as a *Fourier transform*. However, the Fourier transform of one sequence of characters differs from the transform of other sequences of characters. What we really want is a special average (called an *ensemble average*) that smooths the results of all the possible Fourier transformed bit streams that you could send. In a sense, this kind of ensemble averaged Fourier transform (now called a power spectral density or simply a spectrum) best represents the FSK spectrum of *any* message, because it's the average of the spectra of all possible messages. This spectrum, then, tells you how much power exists in a given bandwidth anywhere near the center frequency of the FSK signal.

There are at least three ways to evaluate the ensemble-averaged Fourier transform of an FSK signal. The first, a mathematical approach, uses a description of FSK (either with discontinuous or with continuous phase at the bit transitions), to calculate

*Strictly speaking, there's really only one parameter in FSK, i.e. the phase change per bit, but Amateurs usually speak of frequency shift and wpm or baud rates. With respect to the center radian frequency $(\omega_1+\omega_2)/2$, the phase change per bit ψ is expressed in radians as $\psi = T\times$ shift $\times \pi$ where T is the data bit length in seconds, and the shift is in Hertz.

Table 1. Equation for power spectral density of discontinuous-phase - frequency-shift-keying.

$$DPFSK\ (\omega) = \frac{\sin^2\left[\frac{(\omega - \omega 1)T}{2}\right]}{(\omega - \omega 1)^2} + \frac{\sin^2\left[\frac{(\omega + \omega 1)T}{2}\right]}{(\omega + \omega 1)^2} + \frac{\sin^2\left[\frac{(\omega - \omega 2)T}{2}\right]}{(\omega - \omega 2)^2} + \frac{\sin^2\left[\frac{(\omega + \omega 2)T}{2}\right]}{(\omega + \omega 2)^2}$$

where:

ω = radian frequency
$\omega 1 = 2\pi f 1$
$\omega 2 = 2\pi f 2$
T = the data bit time or the reciprocal of the baud rate

Table 2. Equation for continuous phase - frequency-shift-keying.

$$CPFSK\ (\omega) = \frac{2\sin^2\left[\left(\frac{\omega - \omega 1}{2}\right)T\right]\sin^2\left[\left(\frac{\omega - \omega 2}{2}\right)T\right]}{T[1 - 2\cos[(\omega - \alpha)T]\cos\beta T + \cos^2\beta T]} \cdot \left[\frac{1}{\omega - \omega 1} - \frac{1}{\omega - \omega 2}\right]^2$$

$$+ \frac{2\sin^2\left[\left(\frac{\omega + \omega 1}{2}\right)T\right]\sin^2\left[\left(\frac{\omega + \omega 2}{2}\right)T\right]}{T[1 - 2\cos[(\omega + \alpha)T]\cos\beta T + \cos^2\beta T]} \cdot \left[\frac{1}{\omega + \omega 1} - \frac{1}{\omega + \omega 2}\right]^2$$

where $\omega 1$, $\omega 2$, and T are as in the DPFSK case, and $\alpha = \frac{\omega 2 + \omega 1}{2}$ $\beta = \frac{\omega 2 - \omega 1}{2}$

the spectral density equation.[3] The techniques are powerful and the equations are easily plotted and evaluated.

The second method for evaluating the spectra of FSK signals is to simulate a time-domain FSK signal on a computer, take its Fourier transform using a well-known algorithm, and repeat the process several times using different bit streams. The results are ensemble-averaged so that a spectral density begins to emerge from all the individual Fourier transforms. This method requires considerable computer memory and must be repeated every time you want to try a new bit rate or shift.

The third method for obtaining FSK spectral densities is to use an FSK modulator and spectrum analyzer. However, once you use the equations and come up with a shift and baud rate that you like, you can borrow a spectrum analyzer and see how well your theory meets with practice. (See the two examples at the end of the article).

mathematical analysis

The spectral densities of DPFSK and CPFSK have been calculated by an exact mathematical analysis. These results are then easily plotted on a home computer, in fact, we'll give some BASIC listings to help you along. The equation for the power spectral density of discontinuous-phase frequency-shift-keying is

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shown in **table 1**. The equation for continuous-phase FSK is shown in **table 2**.

Frequency shift is simply $f2 - f1$ *(Hz)*, or $\omega 2 - \omega 1$ in radians per second. Note that if your frequency shift is in a whole number of Hertz, then $\omega 2 - \omega 1$ is never an integer.

T, the data bit length, can be found for Baudot or ASCII data rates and speeds in any recent edition of the *ARRL Handbook*. For the ASCII case, the data pulses and stop pulses are the same length ($T = 3.333$ mS) for the data and stop bit lengths with 300 baud ASCII, so the spectral density results are exact. However with Baudot, the stop pulse is often 50 percent longer then the data pulses. Since there are five data pulses for each stop bit, and the longer stop pulse would only result in a narrower spectrum than the faster data pulses, I use the data pulse time for the following plots. For example, 60 wpm Baudot (baud rate is 45.45) had $T = 22$ mS. This should be sufficiently accurate for any practical purpose.

An Atari BASIC program listing for both these equations is provided in **fig. 3**. The program calculates either the DPFSK spectrum (use **lines 60** to **120** and delete **130-280**) or the CPFSK spectrum (use **lines 130-280** and delete **lines 60** to **120**). The result is placed in array Y (in dB), with the frequency (in Hz) in the X array. The shift and bit times can be changed in **line 20**; they are shown for 170 Hz shift, 60 wpm Baudot. If you link this program fragment with your machine-specific plotting package, you can generate plots like the ones that follow. Atari BASIC is sufficiently generic so that the listing should run on nearly any personal computer with few changes.

spectral plots

The spectral density plots, **figs. 4** through **13,**, are for two cases: the FSK signal with discontinuous phase, and the same signal with continuous phase at the bit transitions. The magnitudes are relative and are intended for comparison purposes only — for instance, to show how sidelobe levels compare from one spectrum to another. It may be puzzling to some that there are no lines in these plots, no periodic signals or carriers, particularly at $f1$ and $f2$. Recall, though, that this is a frequency-modulated waveform with a random bit stream as its input, and in general this kind of modulation result is in a purely broad spectrum with no carriers. The only exceptions occur when special relationships between $f1, f2$, and T exist, e.g. $\omega 2 - \omega 1$, $\omega 2 \div \omega 1$ are integers or 2αT or 2βT are odd multiples of $\pi/2$. For these so-called degenerative cases, the reader is referred to the original Bell Labs paper.[3]

Note particularly how, in each case, the continuous-phase FSK spectrum rolls off much faster than its discontinuous phase counterpart; in fact, this is a characteristic of all continuous-phase signals. This means

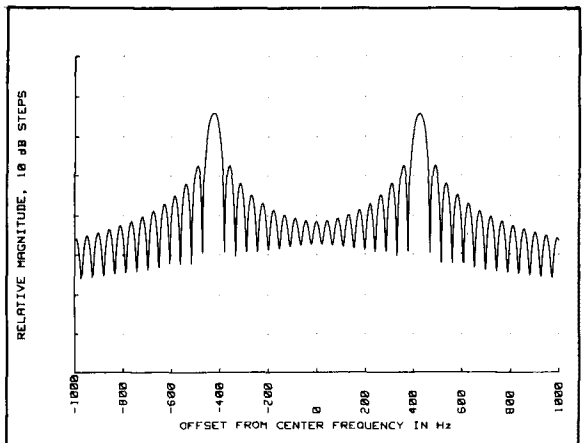


fig. 4. FSK spectrum, discontinuous phase 850 Hz shift for 60 WPM Baudot RTTY.

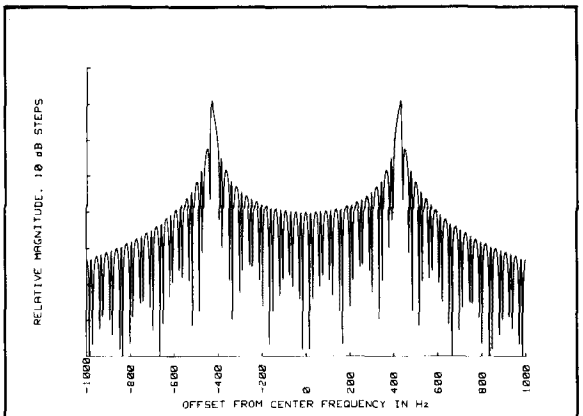


fig. 5. FSK spectrum, continuous phase 805 Hz shift for 60 WPM Baudot RTTY.

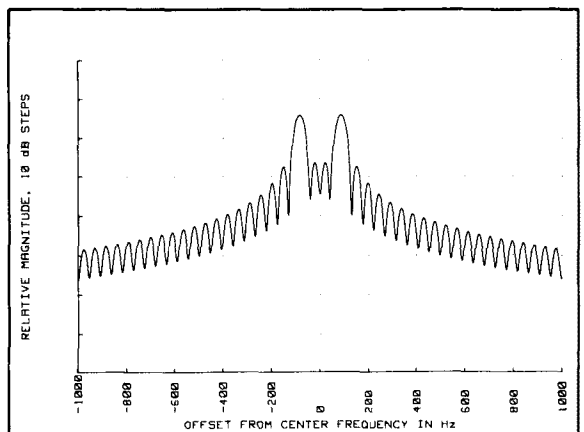
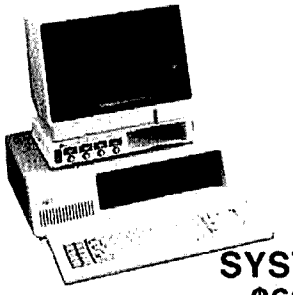


fig. 6. FSK spectrum, discontinuous phase 170 Hz shift for 60WPM Baudot RTTY.

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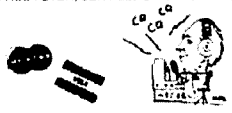
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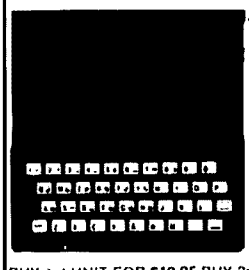
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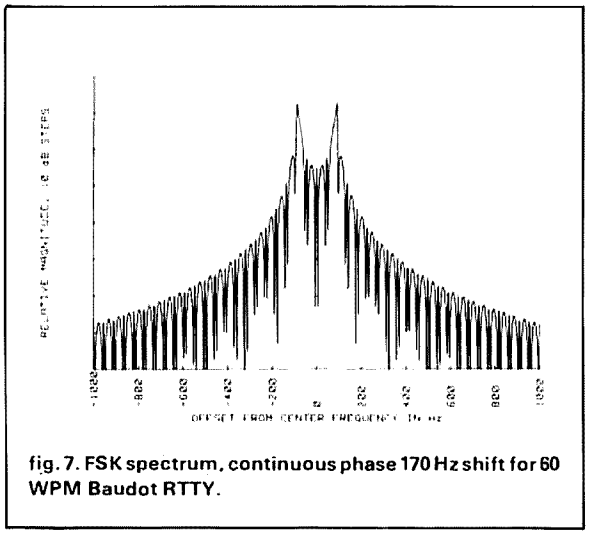


fig. 7. FSK spectrum, continuous phase 170 Hz shift for 60 WPM Baudot RTTY.

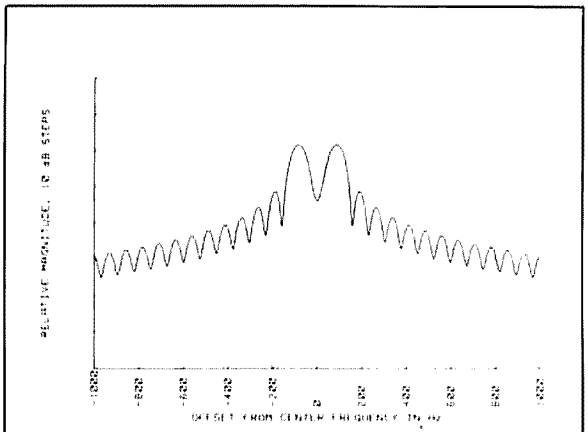


fig. 8. FSK spectrum, discontinuous phase 170 Hz shift for 100 WPM Baudot RTTY.

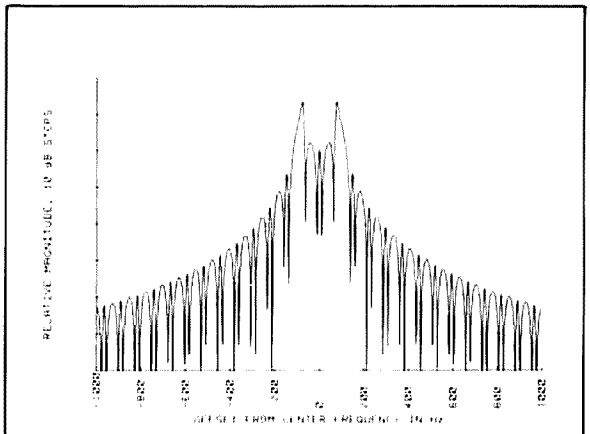


fig. 9. FSK spectrum, continuous phase 170 Hz shift for 100 WPM Baudot RTTY.

that more energy is concentrated near the center frequency of a CPFSK signal, with less energy further from the center frequency that might cause interference with another band user. In short, CPFSK signals are generally more spectrally efficient than DPFSK ones, though spectral efficiency is subject to many definitions. This CPFSK advantage is very evident in the 300 baud, 170 Hz shift ASCII case, where the DPFSK signal is splattered all over the band, and the CPFSK spectrum begins to look like an ideal flat-topped bandpass filter! We'll return to this case shortly.

cumulative power

One advantage of having the mathematical expression for the spectral density is that operations such as integration or differentiation can be performed that would be less accurate and more time-consuming using any other method. Using the expressions given above, the cumulative power in any FSK spectral density can be calculated. By "cumulative power," we mean the following: imagine a frequency window, located at the center frequency of the spectrum, and ever-widening. Each time you widen the window a little, you stop and ask how much of the total power in the spectrum is outside your window's frequency limits. For example, when the window is very narrow, nearly all the spectrum's power is outside. As you widen it, less power is outside the window and your answer gets smaller and smaller. This is then an excellent measure of how well your spectrum concentrates its power in a given bandwidth. It can, for example, be used to estimate the interference levels a given number of Hertz from your signal, or even to design an optimum receiver filter that will capture as much signal power as possible without sacrificing adjacent channel interference rejection.

The cumulative power is obtained by integrating the power spectral densities obtained above. This was not done mathematically, but instead numerically on a computer using the short Atari BASIC listing given in **fig. 14**. The idea is to evaluate the spectral density at ever-widening frequency points, each time adding up the result and scaling by the frequency increment. Numerical integration can be a subtle and tricky business, especially in computer-aided circuit analysis, but these spectra are nicely behaved and the algorithm shown works fine. Note that you use either **lines 110** and **120** (DPFSK) or **lines 140** and **150** (CPFSK); the shift and bit times are given in **line 60** (here 170 Hz shift, 60 wpm Baudot). The results (in dB) are placed in the Y array, and the window width in Hertz is in the X array, ready to pass to your plotting routine. Finally, notice how simple the formulas for the spectral densities become when compared to those in **fig. 3**; this is the result of some clever substitutions made

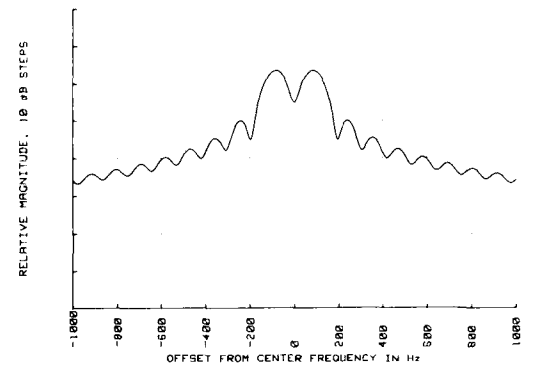


fig. 10. FSK spectrum, discontinuous phase 170 Hz shift for 110 Baud ASCII RTTY.

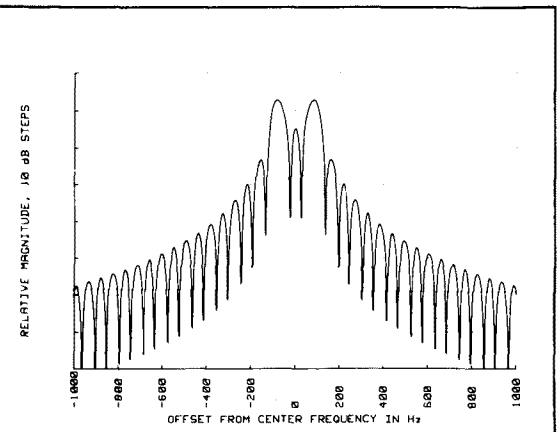


fig. 11. FSK spectrum, continuous phase 170 Hz shift for 110 Baud ASCII RTTY.

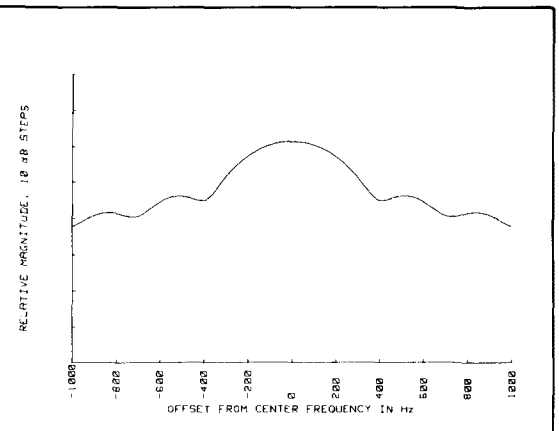


fig. 12. FSK spectrum, discontinuous phase 170 Hz shift for 300 Baud ASCII RTTY.

by sliding the spectra down to a center frequency of 0 Hertz (DC).[4]

The plots given in **figs 15** and **16** compare the DPFSK and CPFSK cases of cumulative power for a Baudot and ASCII waveform. As one would expect from the previous plots, the curves with the faster roll-off are for CPFSK spectra. Note how just a few hundred Hertz away from the center frequency, the CPFSK signal has over 20 dB less power in it than the DPFSK one, an important consideration for evaluating adjacent channel inteference. Other shifts and speeds can be explored using the listing in **fig. 14**.

applications

There are many additional issues one could raise concerning Amateur FSK transmission. One example is the receiver/demodulator design. It was suggested that the receiver filter could be optimized for minimum signal power loss and maximum adjacent channel interference rejection by using the cumulative power plots for a particular FSK waveform. However, the demodulator also has filters in it; the classical Amateur design uses two narrow bandwidth filters centered on *f1* and *f2*, respectively. This type of demodulator filtering makes good sense when receiving FSK transmissions of the kind illustrated in **figs 4-11**, because the filters match the power peaks in the signals, also centered on *f1* and *f2*. However, the 170 Hz shift 300 baud ASCII FSK in **figs. 12 and 13** are another story; here the signal energy is very broad and peaked at the center frequency of the signal, not at *f1* and *f2*. Hence, as illustrated in **fig. 17**, the conventional Amateur FSK demodulator filters waste signal power; their sharply peaked *f1* and *f2* responses no longer match the spectral density. That many Amateurs shy away from 300 baud ASCII on the air begins to make sense . . . their non-optimal modems ignore so much signal energy that they experience high error rates under the usual noisy band conditions. Solutions could involve anything from a radically different 300 baud demodulator design (using a signal correlator, perhaps) to the simple expedient of changing the frequency shift to a new value large enough so that *f1* and *f2* peaks reappear in the spectrum. Of course this will take up more bandwidth, and the former filters will have to be retuned for the new shift, but a definite improvement would be noted over present 300 baud ASCII error rates.

Another issue raised in our spectral analysis of Amateur FSK is its application to spread spectrum communications. The "direct spreading" approach disperses the signal energy so broadly that it lies below the noise level; the signal hides in the mush and goes undetected by conventional narrowband receivers. The spread spectrum receiver is constructed so as to reconstitute the signal from all those bits of energy strewn here and there over the band. The question is, which waveform spreads the energy best? If you look at **fig. 13**, the 170 Hz shift, 300 baud ASCII

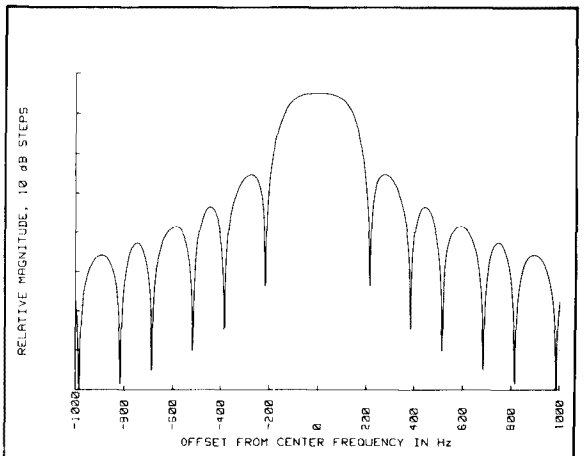


fig. 13. FSK spectrum, continuous phase 170 Hz shift for 300 Baud ASCII RTTY.

```
5 REM *** PROGRAM FRAGMENT FOR
 CALCULATING CUMULATIVE SPECTRA ***
10 DIM X(160),Y(160)
20 PI=3.14159265
30 INC=PI/60
40 NJ=480
50 SUM=0:PT=0:NZ=3:Z=0
60 SHIFT=170:T=0.022
70 A=PI*SHIFT*T
80 FOR J=1 TO NJ
90 B=(J-0.5)*INC
100 REM **** DISCONTINUOUS PHASE CASE
--- USE LINES 110 AND 120 ****
110 YF=2*SIN((B+A)/2)*SIN((B+A)/2)/((B+A
)*(B+A))
120 YF=YF+2*SIN((B-A)/2)*SIN((B-A)/2)/((
B-A)*(B-A))
130 REM **** CONTINUOUS PHASE CASE
--- USE LINES 140 AND 150 ****
140 YF=2*A*(COS(A)-COS(B))/(A*A-B*B)
150 YF=YF*YF/(1+COS(A)*(COS(A)-2*COS(B))
)
160 SUM=SUM+YF*INC/PI
170 Z=Z+1
180 IF Z=NZ THEN YF=1-SUM:PT=PT+1:Y(PT)=
10*CLOG(YF):Z=0
190 NEXT J
200 FOR L=1 TO NJ/3
210 X(L)=L*NZ*INC/(PI*T)
220 NEXT L
```

fig. 14. Atari BASIC listing for calculating the cumulative spectra of FSK signals. The trig functions evaluate their arguments in radians (not degrees).

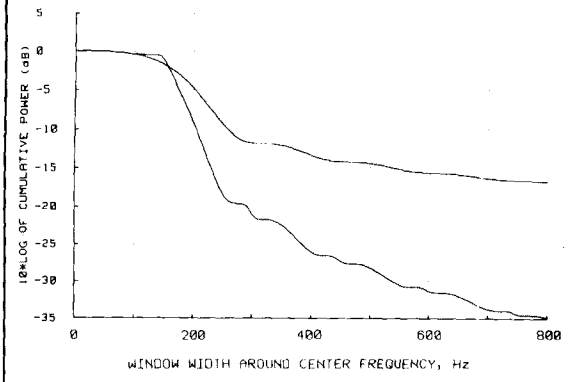


fig. 15. Cumulative spectra of discontinuous and continuous phase FSK for 100 WPM Baudot RTTY, 170 Hz shift.

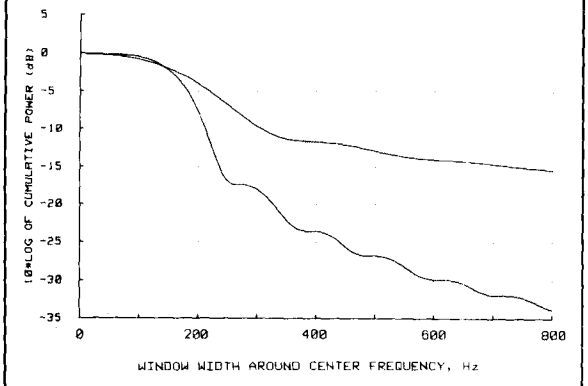


fig. 16. Cumulative spectra of discontinuous and continuous phase FSK for 110 Baud ASCII RTTY, 170 Hz shift.

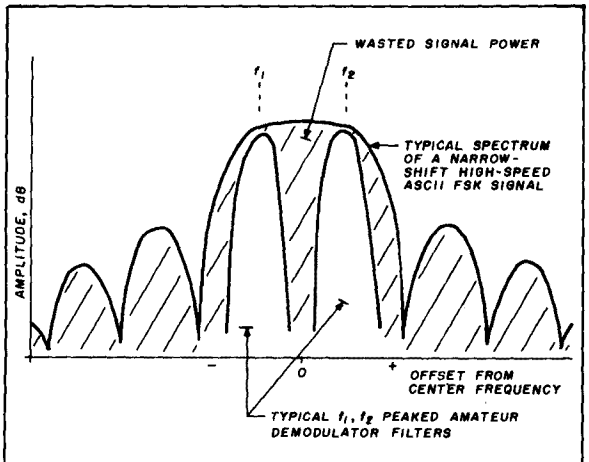


fig. 17. A typical narrow-shift high-speed ASCII FSK spectrum showing how the usual peaked Amateur modem filters ignore a significant fraction of the signal energy.

CPFSK spectrum, you'll begin to get a hint. It can be shown that if you want to fill a given channel or bandwidth with signal energy, the most efficient way to do it is with a rectangular spectral density, because every chunk of available spectrum space will be filled with an equal amount of signal power. Coincidentally, a flat rectangular spectrum also looks most noiselike over the given channel (the spectrum of noise is flat). Now turn to **fig. 18**. It's still 300 baud CPFSK ASCII, but we've changed the shift to 190 Hz. Most striking is the nearly rectangular main lobe, which ideally meets the requirements outlined above. Of course, if you really want to spread your signal energy around and hide the communications under the noise, you can scale everything up by any number you choose; although a 190-Hz shift, 300 baud signal has the same spectral shape as a 19-MHz shift, 30 Megabaud CPFSK signal, it's 100,000 times wider. You could, with this CPFSK waveform and the proper hardware, carry on communication over the entire 420-450 MHz Amateur band and no one except your intended receiver would know you were there!

To demonstrate that these spectral density ideas really work, I generated that rectangular main lobe CPFSK signal and used a sophisticated spectrum analyzer to both ensemble- average and record the resulting spectral density. Shown in **fig. 19**, it nicely confirms the theory of **fig. 18**. The slight raggedness on the spectrum comes from my not averaging every possible bit pattern, which would have taken an infinite amount of time! But this CPFSK spectrum, and all the ones shown here or that you may choose to investigate on your own, can indeed be generated in the laboratory — even the one with that remarkable flat top.

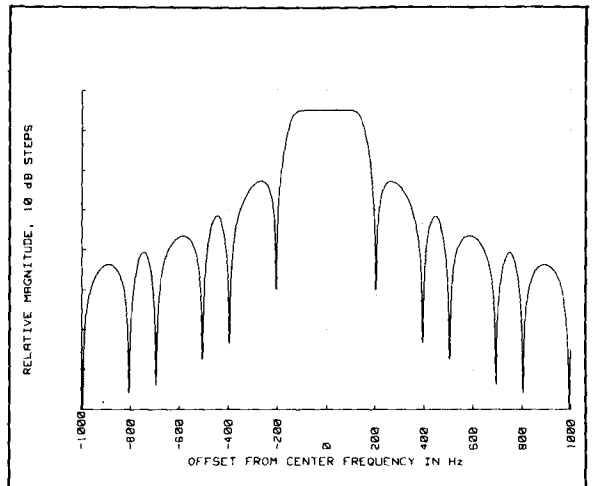
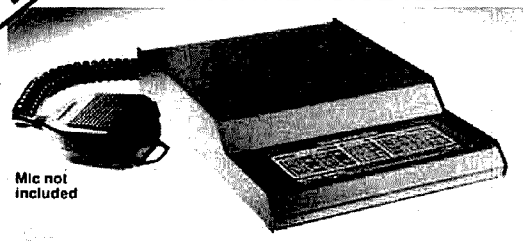


fig. 18. FSK spectrum, continuous phase 190 Hz shift for 300 Baud ASCII RTTY.

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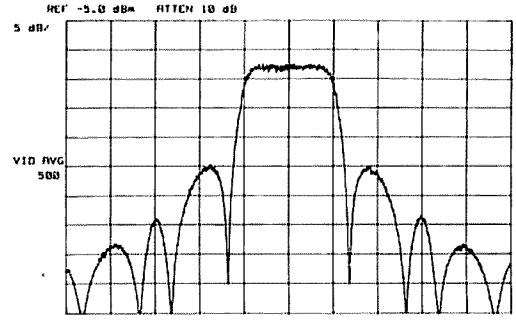


fig. 19. Continuous phase FSK, 190 Hz shift 300 Baud ASCII equivalent.

suggestions for FSK systems

We can summarize this investigation of Amateur FSK spectra with a few suggestions and an invitation. One suggestion is that all Amateurs use continuous-phase FSK in their transmitter modulators. As amply demonstrated above, CPFSK causes much less interference and occupies much less bandwidth than the discontinuous-phase waveform. Fortunately, many modern audio-shift modems already provide phase-continuous signals. Another suggestion is that Amateurs work on FSK demodulator designs which better match the spectral properties of the signal they're trying to receive. This will become increasingly important as bit rates continue to rise, whether in ASCII or in packet communications. Any spectral mismatch in demodulator filtering simply wastes the available (and precious) signal power.

A final suggestion is that Amateurs interested in spread spectrum communications look closely at the family of FSK signals as good candidates for a direct spreading waveform. System performance could significantly improve once the waveform's spectrum is matched to the spreading requirements.

Much more could be said about Amateur FSK — in receiver and transmitter design, in error rate performance, and in application to new types of communications systems. This simple binary keyed waveform carries with it several elegant and easily explored spectral properties, characteristics outlined in this article. I invite the Amateur community to investigate further and make their own fascinating discoveries and conclusions.

acknowledgements

Thanks to Dr. E.J.Kelly, L.A. Bessette, and R.F. Giovannucci for their original contributions to the thesis upon which this article was based.

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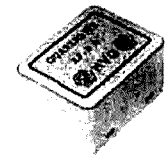
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ham radio TECHNIQUES

Bill Orr W6SAI

MacGyver of the '30s— how Amateur ingenuity saved the day

In many man-made and natural emergencies Amateur Radio has proven itself again and again. Here's the story of how Henry Jenkins, W7DIZ, responded when one of the heaviest storms of the 1930s hit the Oregon coast and blinded the Tillamook Rock lighthouse.

The lighthouse, which protected coastal shipping from the dangerous shoals, was under the control of the Department of Commerce. The light atop its tower was perched 85 feet above the ocean. Power for the light, the radio, and the small dwelling of the lighthouse keeper was derived from a gas generator in the basement.

The year is 1934; the month, October. . . .

About 10 p.m., October 20, a fresh southeast wind was blowing with light rain. During the night, the wind increased to gale force and changed to the southwest. About 3 a.m. the seas ran extremely high, swells hitting the base of the rock from the southwest with the spray coming over the rock.

At 9:30 a.m. I was awakened with a sudden jar, completely covered with water. All my clothes and bedding were soaked. The seas at this time were washing over the entire tower, pounding against the window shutters of my room. The window catch let go and the entire room was flooded. At this time the wind was blowing at an estimated 100 miles per hour, seas covering the lighthouse, carrying with it large rocks, debris, and fish which were smashed through the lantern plate glass — 85 feet above sea level — breaking 16 panes and flooding all quarters. In endeavoring to replace glass panes with emergency wooden panels, the Keeper's right hand was deeply cut. While assisting dressing his wound, I read the barometer, which was 28.92 inches, and dropping. Each time the tons of water would cover the building, coming down with terrific impact on the roof, the barometer would drop immediately to 28.72 inches, returning as soon as the impact was over.

The impact from the water and rocks would occur about every three seconds. About 10:15 a.m. a terrific impact occurred and the tower and building were enveloped with giant seas. The large 80-foot derrick and telephone cable to the mainland had been swept away.

A 6-foot section of the west end of the rock was broken off and carried away, hurling rocks weighing as much as 50 pounds through the tower and to the roof, smashing shutters made of half-inch planking. All floors were flooded, breaking the piping and heating system, thereby cutting off the heat, which was badly needed. We were soaked, sloshing in salt water, and extremely tired. The lantern and fog signal were inoperative. Both of these were greatly needed by mariners, and now we had no means of warning them of the rocks and shoals around the light. Finally, we were able to erect a crude, white light, which we hoped somebody could see. But what was needed was a means of informing the Superintendent on land of the damage and the fact that the light was useless.

What was needed was a radio to communicate to land. Since the power was out (and the government radio equipment had not yet been delivered and installed), it was up to me to do the job. A radio had to be built. All that was at hand was a collection of parts from an old Atwater-Kent broadcast set, a few dry batteries and scraps of tinfoil, copper, and brass.

When the storm abated a bit, W7DIZ went to work: *I got two pieces of relatively dry wood. Not having any tube sockets, I drilled holes in each board to take a tube and with the aid of a gas torch, soldered leads on the tube prongs.*

On the transmitter board I placed a coil which I wound from wire taken from an audio transformer. The coil was wound on the cardboard case taken from a No. 6 dry cell. I used 14 turns. The tuning capacitor was salvaged from the old broadcast receiver. Capacitors were made of tinfoil and wax paper taken from a loaf of bread. Since no resistors were at hand, none were used. An rf choke was made from a broadcast coil by removing 100 turns. This junk, when finally assembled, completed a TNT (tuned-not-tuned) oscillator, but had no grid condenser or grid resistor. The batteries to power the transmitter were taken from the defunct telephone system. Eighty volts was available. For a key, I simply broke the battery connection with my hand. Later, I found a piece of spring brass and made a hand key. The antenna consisted of 40 feet of salvaged wire from a broadcast coil.

The receiver was next. For a coil, an old telephone receiver was used as the form and 45 turns of wire stolen from one of the broadcast coils was wound on the form. This made the grid coil. Twenty turns were added at one end for the plate coil. There was no condenser available to control oscillation,



Photo A. **After the terrible storm, W7DIZ rebuilt his "junk box" radio station. At left is the one-tube transmitter, with the copper tubing tank coil. In the center, in front of the tool box, is the one-tube receiver. A heavy brass hand key is positioned on the table near the transmitter. With this simple, home-made equipment, W7DIZ maintained schedules and kept the Tillamook light on the air, in the ham bands.**

so enough turns were put on the plate coil to keep the circuit oscillating at all times. For the antenna series condenser, two feet of insulated wire were twisted together. A tinfoil condenser was used for the grid condenser, but there was no resistor for the grid leak.

The tuning condenser was made from two brass plates removed from the door knob. One plate was screwed to the wood board and the other placed atop it, separated by a piece of waxed wrapping paper. Tuning was accomplished by shoving one plate over the other with a pencil.

By cut-and-try, I finally heard signals in the 80-meter band. The first station heard was W7RT in Seattle. |Since most stations in those days were crystal-controlled and remained at one spot in the band, W7DIZ immediately knew where he was. — W6SAI|

I next heard W7CKX, a station relatively close to the light. I left the receiver tuned to that spot and tuned the makeshift transmitter to the same spot by listening to CKX with the transmitter running. Finally, about 6:30 p.m. I called W7CXK. No response. But the station CXK was working — W7WR— told CXK that somebody was calling him! And the contact was made. I sent a message to the Superintendent of Lighthouses, advising him that the light was out of service, and that all shipping should be warned. W7CXK immediately sent the message via Western Union to the Superintendent in Seattle and within minutes the powerful Department of Commerce radio stations up and down the Pacific Coast flashed the news that the Tillamook light was inoperative.

And that's how Henry Jenkins, W7DIZ, succeeded in building an entire radio station from salvaged pieces of junk, put the rig into operation, and effected communication with other Amateur stations in an emergency of great import to the shipping lines along the western coast of the United States.

Within the next few months the light was repaired and the Lighthouse Service, seeing the need for radio communications at the Tillamook light, provided a new, more powerful generator and radiotelephone equipment. A lighted buoy was placed near the light. Jenkins rebuilt his ham gear (see **Photo A**), still using only a single tube in the transmitter and receiver.

At one time, W7DIZ took a telephone microphone, connected it to the transmitter tank coil via a loop and made the little rig into a phone transmitter! W7CXK said the audio quality was excellent, provided Henry didn't move around very much!

While the new radio equipment was being installed, W7WR and W7CXK were hired by the Lighthouse Service (at 50 cents an hour) to provide communication with the light, using the special calls KCBC-2 and KCBC-3. This lasted about four months, running three schedules a day. Many private messages to the families of the men on the isolated lighthouse were handled and, as W7CXK says, "Many hours were spent trying to keep the men from going stir-crazy on the rock. It is unbelievable what a few months of isolation out there could do to a person's mind."

Eventually, the lighthouse came under the supervision of the Coast Guard. In the mid 1960s, the light was no longer required, so it was sold. But as recently as 1971, the buyers of the light had done nothing to it, and it looked much the same as ever.

MacGyvers of the 80s? — a "make-do" hamfest event

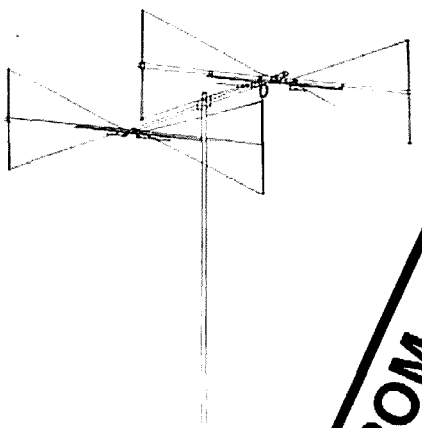
Does the "Make-do" spirit still exist in Amateur Radio? I like to think that it does. In this regard, I want to propose a "Make-do Event" at your next local hamfest or convention. Volunteer hams split up into three or four teams, three hams to a team. At hand is a large junk box full of old parts, perhaps a defunct TV set, and one or two old radios. Each team has a small number of appropiate tools — a screwdriver, wirecutters, and pliers — as well as a small collection of nuts, bolts, *and woodscrews.*

Now the contest begins. It runs for a given period, say, three hours. The goal is to build, literally from junk, radio transmitters and receivers and to effect communication between two groups. The groups that get on the air and establish communication first are, of course, the winners.

Details have to be worked out. Is it permissible to scavenge for wood or other chassis material? I vote *yes*. Can a soldering iron be used? I vote *no*. I

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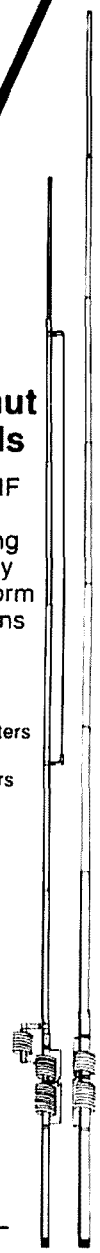
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also vote that the transmitter be crystal-controlled, to keep all activity in the ham bands. Alternatively, a low-power "standard frequency" signal could be emitted by the managers of the contest so that contestants can "find the band."

With a little thought, a viable and exciting contest can be created. I hope such a contest will prove that the spirit of "Make-do" still exists in Amateur Radio today.

The original story of Henry Jenkins, W7DIZ, was told in the January, 1935 issue of "Radio" magazine. Additional details of the incident were given to me in private correspondence from Henry Goetze, W7CXK, who was a witness and participant in the incident at Tillamook Rock.

ham radio

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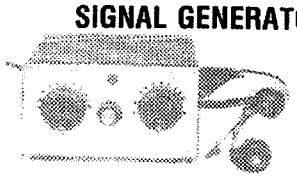
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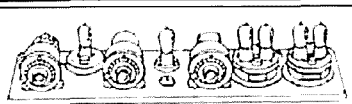
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PRACTICALLY SPEAKING ... Joe Carr K4IPV

"function generator" circuits from your signal generator: part 1

Many readers own sine wave or square wave generators, or can make them using simple circuits. In this article, I'll show how to make a function generator out of simple *audio* signal generators. First, we'll look at making square waves out of sine waves, and then follow up with generating triangle and sawtooth waveforms.

keeping the circuits stable

In preparing this article I had to breadboard several circuits (integrators) to check out the principles presented. Ordinarily, common 741 operational amplifiers are used for my casual projects because those devices are both low-cost and well-behaved. Unfortunately, their "well-behaved" quality is due to their unconditional stability; in other words, the frequency response is rolled off so much that there's little chance of oscillation. That fact makes it easy for the builder to become slack in such matters as power supply bypassing. The 741 will usually operate normally with the power supply leads unbypassed, but op-amps with higher frequency responses (such as the CA-3140 BiMOS device, with its 8-MHz gain-bandwidth product) will become unstable and oscillate if the power supply is unbypassed.

Figure 1 shows the first circuit that I connected. Because the integrator circuits being checked are based on the inverting follower, I first connected an inverting follower with a gain of 1 (R1 = R2 = 470 k) to test the op-amp from my junk box and the breadboard connections. The idea was to then replace the appropriate resistors with the correct integrator components. When a 400-Hz sine wave signal was applied to the input, the oscilloscope showed the waveform in **fig. 2**. Note the thickening of the trace; the fuzziness indicates oscillation at high frequency. Expanding the trace demonstrated that the oscillation was at a frequency somewhat higher than 20 kHz. (**fig. 3**).

The CA-3140 operational amplifier has a higher gain-bandwidth product than the older 741, and requires the bypass capacitors shown in **fig. 4**. When I connected those capacitors into the circuit, the oscilloscope showed the trace in **fig. 5**, which is the correct trace. Beware! Always follow good shop, lab, and design practices.

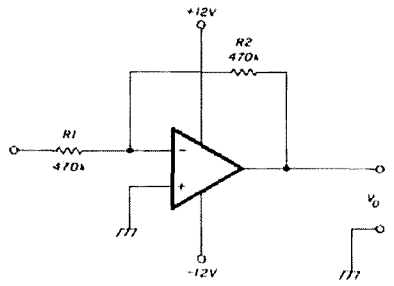


fig. 1. An inverting follower tests the op-amp from junk box and the breadboard connections.

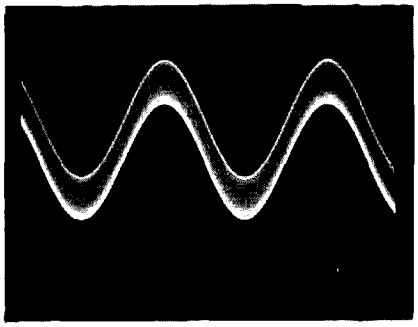


fig. 2. 400-Hz output with superimposed 20-kHz oscillation.

square waves from sine waves

Figure 6 illustrates a method of converting sine waves into square waves. The circuit is an operational amplifier connected as a comparator. Because the op-amp has no negative feedback path, the gain is very high; with typical op-amps gains of from 20,000 to 2,000,000. Thus, a voltage difference across the input terminals of only a few millivolts will saturate the outputs. From this behavior, we can understand the operation of the circuit and the waveform in **fig. 6**.

The input waveform is a sine wave. Because the noninverting input is grounded, the output of the op-amp is zero only when the input signal voltage is also zero. When the sine wave is positive, the output signal will be at $-Vo$; when the sine wave is negative, the output signal will be at $+Vo$. The output signal will be a square wave at the sine wave frequency, with a peak-to-peak amplitude of 2Vo. An output level control, variable gain post-amplifier stage or *an attenuator* network can be used to reduce the signal to a usable level.

Before we turn our attention to circuits for generating other waveforms, let's investigate the topic of operational amplifier integrators.

integrators

Electronic integrators must find the time average of the input waveform. The simplest form of integrator is the resistor capacitor (RC) network shown

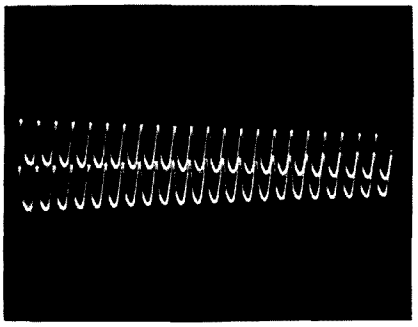


fig. 3. Expanded trace shows oscillation signal.

in **fig. 7.** In this circuit, the output voltage change is related to the capacitor's rate of charge. The rate of increase in Vo depends upon the input voltage and the values of R and C. If you recognize the circuit as a low-pass filter, you are correct. The integrator also functions as an LPF with a roll-off of −6 dB/octave above the −3 dB knee frequency (1/6.28 RC).

An active integrator is shown in **fig. 8.** The active element is an operational amplifier, and it has a capacitor in the feedback network. This circuit is basically the inverting follower with the feedback resistor (R2) replaced by a capacitor. The transfer equation of this circuit is given by the expression shown in **fig. 8.** In this expression, the resistance is in ohms, and the capacitance is in farads. The term "a" is a constant that represents the initial charge on the capacitor, which could be zero. The constant factor −1 indicates that the circuit is inverting, so a positive-going input voltage produces a negative-going output voltage, and vice versa.

The gain of this circuit is −1/RC. Keep in mind that with small values of R and C, this gain can be very, very high, with resulting problems for the designer. Consider this example: what is the gain of an inverting integrator in which R = 10 k and C = 100 pF?

$$GAIN = -1/RC$$
$$GAIN = -1/((10{,}000\ ohms) \times (100 \times 10^{-12}\ farads))$$
$$GAIN = -1/0.000001$$
$$GAIN = -1{,}000{,}000$$

At a gain of 1,000,000 we find that

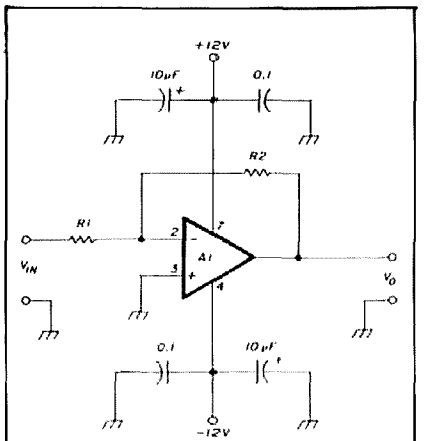


fig. 4. CA-3140 op-amp requires bypass capacitors, as shown.

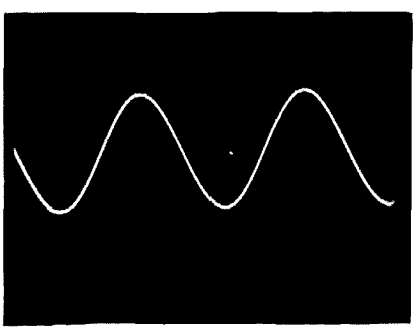


fig. 5. "Cleaned up" waveform from corrected circuit.

with a small error or dc offset in the input signal, the output will be tremendous. For example, suppose we have a 1-volt peak-to-peak sine wave that has a 10-mV dc offset (not very much). This dc offset will be integrated with a gain of 1,000,000 so that 10 mV becomes (10mV) x (1,000,000) = 10,000 volts. Of course, the output of the op-amp is limited to about 10 or 12 volts, so the output will saturate quickly. The normal offset of a 741 will saturate the output of such an integrator so fast you'll think it's shorted! In general, the rule of thumb is to make the time constant of the RC network long (i.e., 5X) relative to the period of the input signal.

Figure 9 shows some common electronic signals at the input and output of an electronic integrator. In each case, the upper trace is the input and the lower trace is the output. For the case of the sine wave shown in **fig. 9**, the output is phase-shifted 90 degrees to become a cosine wave. Thus, the output is in quadrature with respect to the input signal. Although only a few Amateur uses for quadrature signals exist, there are some, so an example is included.

The integrator action on square waves is shown in **fig. 10.** Because the amplitude of a square wave is constant, the integrator output will rise at a constant rate until the square wave drops low again. At that time, the slope of the output waveform changes and starts decreasing. Thus, the integrator makes a triangle waveform out of a square wave. Incidentally, this technique is used in commercial function generators to form triangle waveforms.

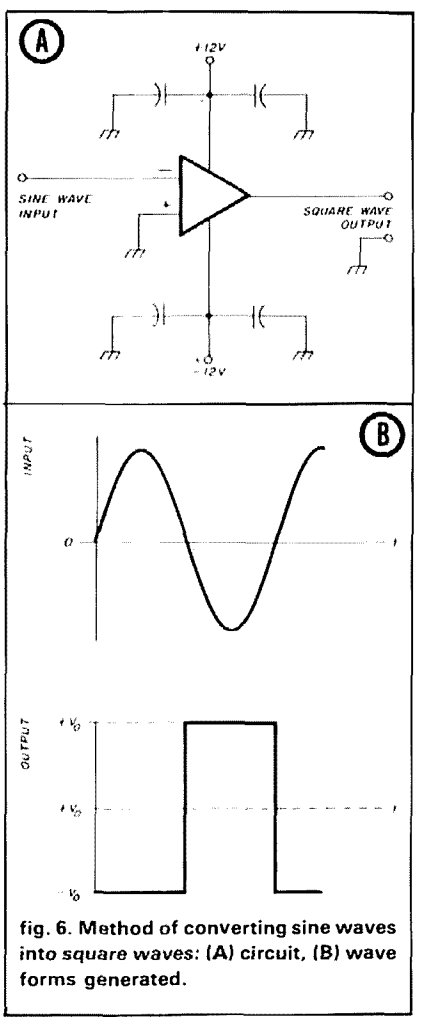


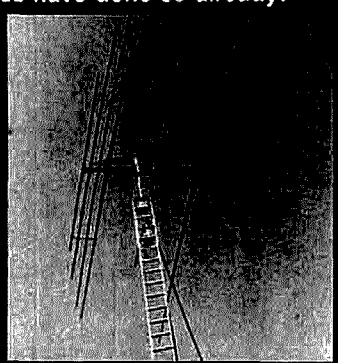
fig. 6. Method of converting sine waves into square waves: (A) circuit, (B) wave forms generated.

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next month

In part 2 we'll examine a practical integrator circuit and some practical sawtooth generator circuits.

more on flood damage

After my column on flood damage appeared in *ham radio* (October, 1985, page 95), I used some of the same material, presented differently, in a radio/TV repair trade magazine. One reader, a former Naval officer who had supervised electronics technicians on shipboard, endorsed my advice and added some of his own.

He described an unusual manner of repairing salt water-soaked equipment. A sailor would take the damaged equipment into the shower and slosh it down thoroughly with warm water (distilled or softened water is preferred, though not always available). He'd then take the desalinated equipment to the galley and dry it out in the ovens with low heat and good air circulation.

The retired officer also suggested cleaning equipment covered with oily dirt with a mixture of 8 to 10 ounces of household ammonia, 4 to 6 ounces of a cleaner such as Mr. Clean™ or Lysol,™* 4 to 6 ounces of acetone (the ingredient in some fingernail polish removers), and enough distilled (or softened) water to make one gallon of solution. Dunk the equipment into this mess. (For larger equipment, proportionally larger amounts can be used.) He also recommended using an old Water Pik™ tooth-cleaning device to hose off equipment too large to dunk; the equipment is then dried in an oven set to 140 to 150 degrees F (note: some plastics used in Amateur equipment will melt at temperatures over 130 degrees, so beware) for 4 to 5 hours. All lubricants in switches, potentiometers, and air-variable capacitors must be replaced after this treatment.

The black asphalt-like goop that oozes out of overheated transformers can be easily removed from chassis by using either freeze spray or a blast from a CO_2 fire extinguisher. (Use one that needs refilling; don't waste your fire protection on cleaning jobs.) The brittle, frozen goop can be flaked off using a dental tool or soldering tool.

ham radio

*Mixing ammonia with any cleaning product (such as chlorine bleach) that contains sodium hypochlorite produces chlorine gas. Always check the label of any household cleaner before combining with ammonia, and be sure ventilation is adequate.

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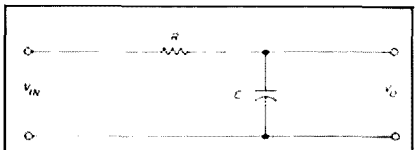


fig. 7. Simplest form of integrator uses two components — a capacitor and a resistor.

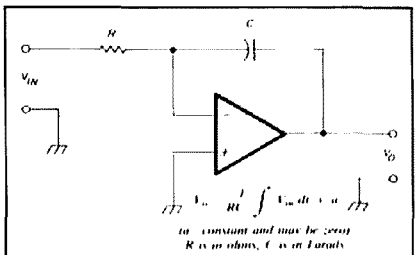


fig. 8. An active integrator uses gain stage plus capacitor and resistor.

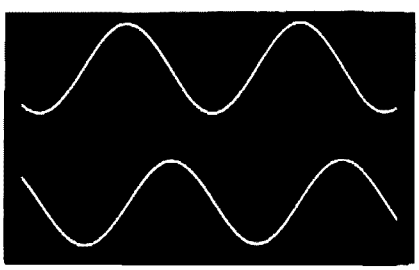


fig. 9. Common input/output signals integrated—sine wave becomes cosine wave (90-degree phase shift).

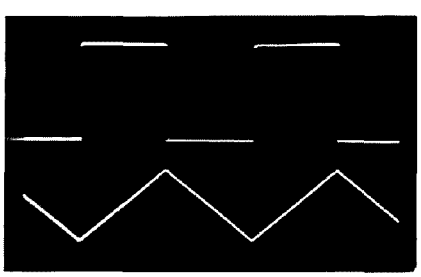


fig. 10. Square waves become triangular waves.

VHF/UHF WORLD

Joe Reisert W1JR

a high-gain 70-cm Yagi

The Yagi is surely the king of antennas below 450 MHz. There have been tremendous strides in Yagi design in the past decade; they include, but are not limited to, increased gain per unit of boom length, very long boomlength designs (5 to 20 wavelengths), improved sidelobe patterns, wide bandwidth, and improved front-to-back ratios.

Last May's VHF/UHF World summarized many of the facts and fallacies about Yagi design. It also described some of the history associated with Yagi design and presented data on approximate gain and boom correction factors. That column[1] should be reviewed before reading this month's.

That column also mentioned using computers to optimize Yagi designs. Understandably, many Amateurs have neither the specific expertise nor the computer capabilities needed to design their own Yagis. These Amateurs prefer either to purchase commercial antennas or build a tried and proven design. The latter is the subject of this month's column.

metamorphosis of a 70-cm long Yagi

In the mid-1970s I became interested in designing a long Yagi to replace my extended expanded collinear.[2] The NBS 4.2-wavelength Yagi was used on tropo and EME, but more gain per antenna was still desired.[3] Proven designs with high gain were lacking and I didn't have the time to "roll my own" design.

In 1977, Günter Hoch, DL6WU, published an article on long boom Yagis, but I didn't give it much thought at the time.[4] However, in 1981, determined to beat the NBS designs, I decided to see if Hoch's claims were valid. The day before leaving for Sioux Falls, South Dakota, to attend the Central States VHF Conference, I built and partially tested a hurriedly-designed 28-element, 21-foot boom, 432-MHz model.

The reaction of other participants in the antenna measuring contest was as expected: "What's W1JR trying to prove now?" "Will it work?" "What a monster!" But it pinned the test meter in the antenna measuring contest. Tension mounted as this design beat all other entrants by almost 2 dB! The topic of conversation quickly changed from "Will it work?" to "How can I get a copy of the dimensions?"

In this particular design, documented in reference 5, elements were placed through the boom using ohmic contact with NBS boom corrections.[3] The boom was made from a 21-foot length of 3/4-inch square aluminum left over from a previous project. Several Amateurs quickly duplicated the design with great success. Other Amateurs used four of these Yagis successfully on 70-cm EME. Performance seemed to equal that of eight of the then-popular shorter boomlength designs, but with far less feed system loss and complexity.

When DL6WU updated his design,[6] I decided to use the new information to build an improved model, particularly in terms of improving the mechanical design. I wanted insulated through-the-boom elements because they're easier to work with and don't require set screws to get good ohmic contact to the boom.

With the help of the new electrical design information, I built a 21- and a 24-foot boom model using a 1-inch square boom. The boom corrections for insulated through-the-boom mountings had been determined earlier on a similar antenna design. The larger boom was much stronger than the previous design and easier to build because it didn't require all the drilling and tapping.[7]

These new designs had higher gain and an even cleaner pattern than the previous models. Closer observation showed that in addition to slightly higher gain on the 24-foot model (which was as expected), this model also had a better front-to-back ratio. Had the sought-after long Yagi antenna design at last been found?

further development

Just about this time computer-aided Yagi design programs became available.[9] Stan Jaffin, WB3BGU, and Steve Powlishen, K1FO, wasted no time in trying to see if this Yagi had all the gain seen on the antenna ranges. It did, but they noticed that the sidelobes, while as good as other moderate-to-long Yagi designs, weren't as good as required for EME. The antenna also appeared to be tuned higher in frequency than desired. Additional range testing seemed to show that the sidelobes, while not down as far as desired (16 to 18 dB below the main beam), were also not as poor as the antenna program indicated.

Then the MININEC program was obtained and tried by several antenna designers. Voilà! Using this program and modeling with eight segments for each element, the computer generated patterns that were amazingly close to those measured in actual tests.

The pattern of the original 24-foot

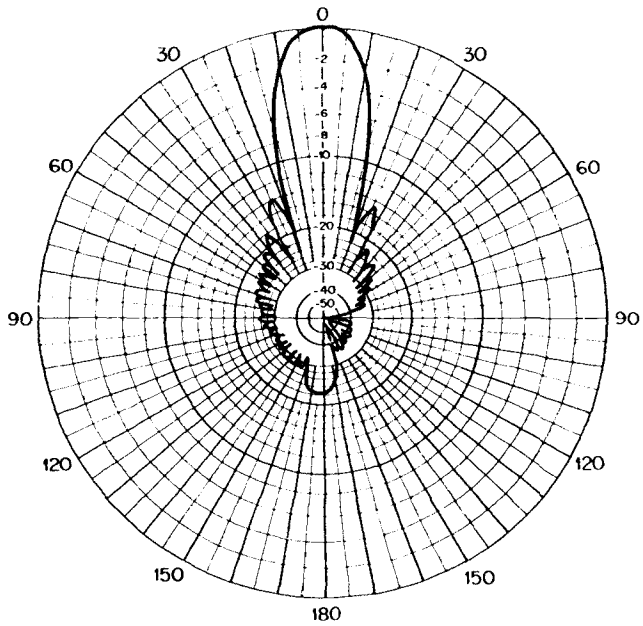


fig. 1. The computed radiation pattern of the original 24-foot boom, 432-MHz Yagi. Note that the E (horizontal) plane is on the right-hand side of the plot and that its first sidelobe is approximately 15 dB below the main beam. The H (vertical) plane is on the left-hand side of the plot and its first sidelobe is approximately 14 dB below the main beam. Gain is approximately 19.15 dBi.

Yagi shown in **fig. 1**, using data from MININEC, seems to very accurately predict the pattern and gain measured on an antenna range. Gain was calculated to be approximately 19.15 dBi with the first sidelobes at 15 and 14 dB in the E and H planes, respectively. Gain on the test range was just above 19 dBi or very close to the predicted value.

Although I still wasn't happy with the pattern, I didn't have a suitable program or computer with which I could try to improve the performance. Because he was designing a new EME array for NC1I,[10] Steve Powlishen, K1FO, was quite interested in the problem. He tried one of his automatic optimization routines on my 31-element design, and after many hours of computer time, found that the first and the last few directors were too short. When these directors were lengthened slightly, the gain increased about 0.25 dB, but the pattern improved considerably with the sidelobes in the E and H plane at approximately 18 and 17 dB, respectively.

Just prior to the 12th annual Eastern VHF/UHF Conference in May 1986, I built the new model. At the antenna measuring party at this conference, the gain was indeed near the predicted value, 19.4 dBi. However, what was even more comforting was that the first sidelobes in the E plane were nearly 18 dB down, right where the computer had predicted they would be. At last, the "almost perfect" long Yagi was working near the theoretically highest possible gain.[1]

I obtained my own version of MININEC and now have it running on a large computer. The plots computed for the optimized 24-foot boom Yagi seem to match the measured pattern closely. Because I can't measure H-plane radiation patterns conveniently, it's quite valuable to have the computer do the pattern work for me. **Figure 2** shows the computer-predicted patterns for the optimized Yagi.

other considerations

Recent tests by many Yagi builders have debunked the theory that long Yagi antenna designs are very critical.[1] In fact, if there is a sufficient number of directors and the original design is duplicated correctly, the opposite may be true. The 1-dB bandwidth may be very large, exceeding 1.5 to 2 percent.

Furthermore, weather conditions such as rain, ice, and snow tend to detune the frequency downwards because the elements become larger in diameter. During extreme icing conditions in past years, I've actually seen the gain of some of the popular Yagi antennas drop almost to zero dB while the antenna actually had more gain off the rear than off the front!

Therefore, it's best to design a Yagi to a slightly higher frequency than the desired operating frequency; 0.7 to 1 percent (3 to 4 MHz at 432 MHz) seems to be a good compromise. Also remember that most Yagi antennas show a gain characteristic similar to that of a lowpass filter. That is, they tend to roll off more slowly in gain as frequency decreases but roll off very rapidly above the design frequency. If the elements are a little shorter than optimum, the gain will be decreased by only a few tenths of a dB at most and the pattern may actually improve at the operating frequency.

high-gain Yagi construction

Figure 3 shows the mechanical details of my latest high-gain 432 MHz Yagi described above. It was originally patterned after the DL6WU model using his recommended element spacings.[6] However, the element lengths shown have been computer-optimized by K1FO. The computer pattern is as shown in **fig. 2**. Four of these Yagis will do very well on EME and eight will be approximately equivalent to 16 of the typical 12- to 15-foot models presently in use.

The boom uses 1-inch square stock with 0.062-inch wall thickness since this is easier to drill and has the equivalent strength of much larger diameter round tubing. Round tubing 1 inch in diameter can be substituted if necessary, but it will be more difficult to drill properly in a home workshop and its

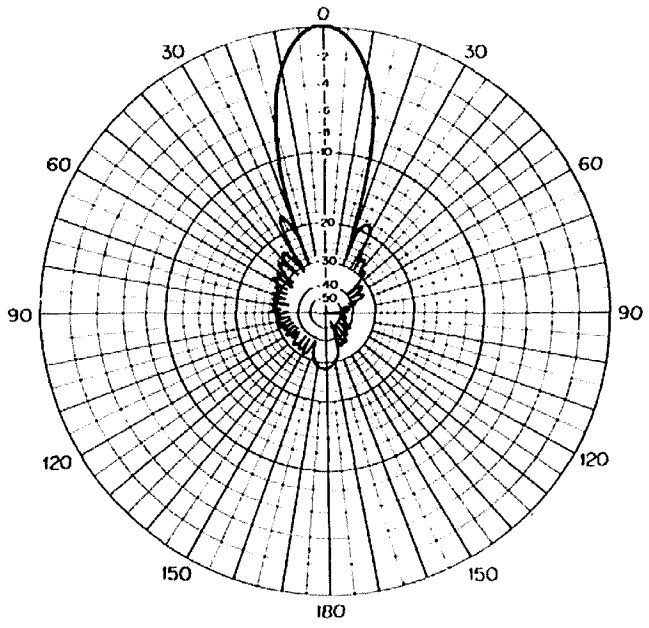


fig. 2. The radiation pattern of the computer-enhanced 24-foot boom, 432-MHz Yagi. Note that the E plane is on the right-hand side of the plot and that its first sidelobe is approximately 18 dB below the main beam. The H plane is on the left-hand side of the plot and its first sidelobe is approximately 17 dB below the main beam. Also note that the *magnitude of* the sidelobes *on* both the E and H planes are 3 dB lower than those of the original Yagi shown in *fig. 1.* Gain is slightly higher at approximately 19.4 dBi.

strength will be less than that of a square boom of comparable diameter.

The boom was chosen to be 24 feet long (10.5 wavelengths) because at this length the DL6WU Yagis have good gain and a naturally high front-to-back ratio, as discussed in reference 1. Also, 24 feet is a standard length of square tubing in the United States and Canada. (21 feet, 1 inch is also a standard length in the United States and can be used by joining with an additional 3-foot section if desired.)

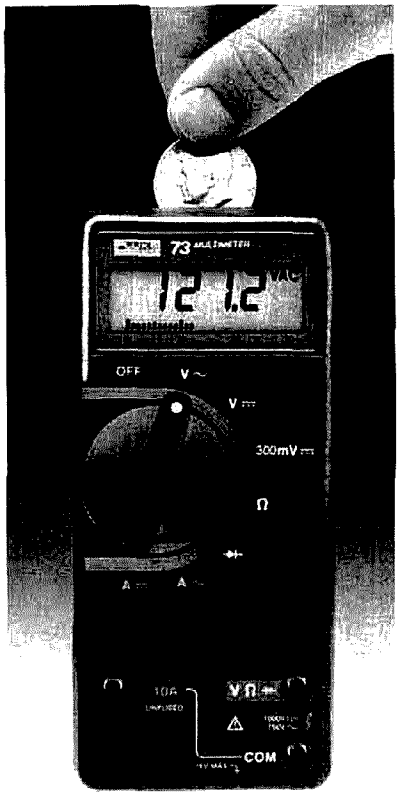
The elements are all made from 3/16-inch diameter aluminum rod and are insulated from the boom as discussed earlier. Elements should have about a 0.032-inch bevel (easily accomplished with a file) at the end of the elements instead of a straight cut. Also, all elements should be held in place with "keepers." If the electrical performance is to be retained, it's important to use this mounting method because this design has a built-in boom correction factor unique to this particular model. Some Amateurs have used teflon inserts that don't require the keepers.[11] Although these are satisfactory, they may require a different boom correction factor that I haven't yet determined.

K9KFR and K3IPW* can each supply the proper insulators and galvanized keepers. Galvanized keepers are adequate, but tend to rust out after extended exposure to moisture. Despite the increased cost, I prefer stainless steel keepers for long-term performance. The model 6100-18 retaining ring from Industrial Retaining Ring Company is the type I use, but other suppliers may have suitable equivalents.**

The feed system uses a "T" match because it's well balanced, easy to adjust, and very efficient. A 4:1 balun using semi-rigid 0.141-type coax,[12] patterned after one of my previous

*Bob Johnson, K9KFR, RR4, Road 600N, Columbia City, Indiana 46725; Tom Rutland, K3IPW, 1703 Warren Street, New Cumberland, Pennsylvania 17070.
**Industrial Retaining Ring Company, 57 Cordier Street, Irvington, New Jersey 07111.

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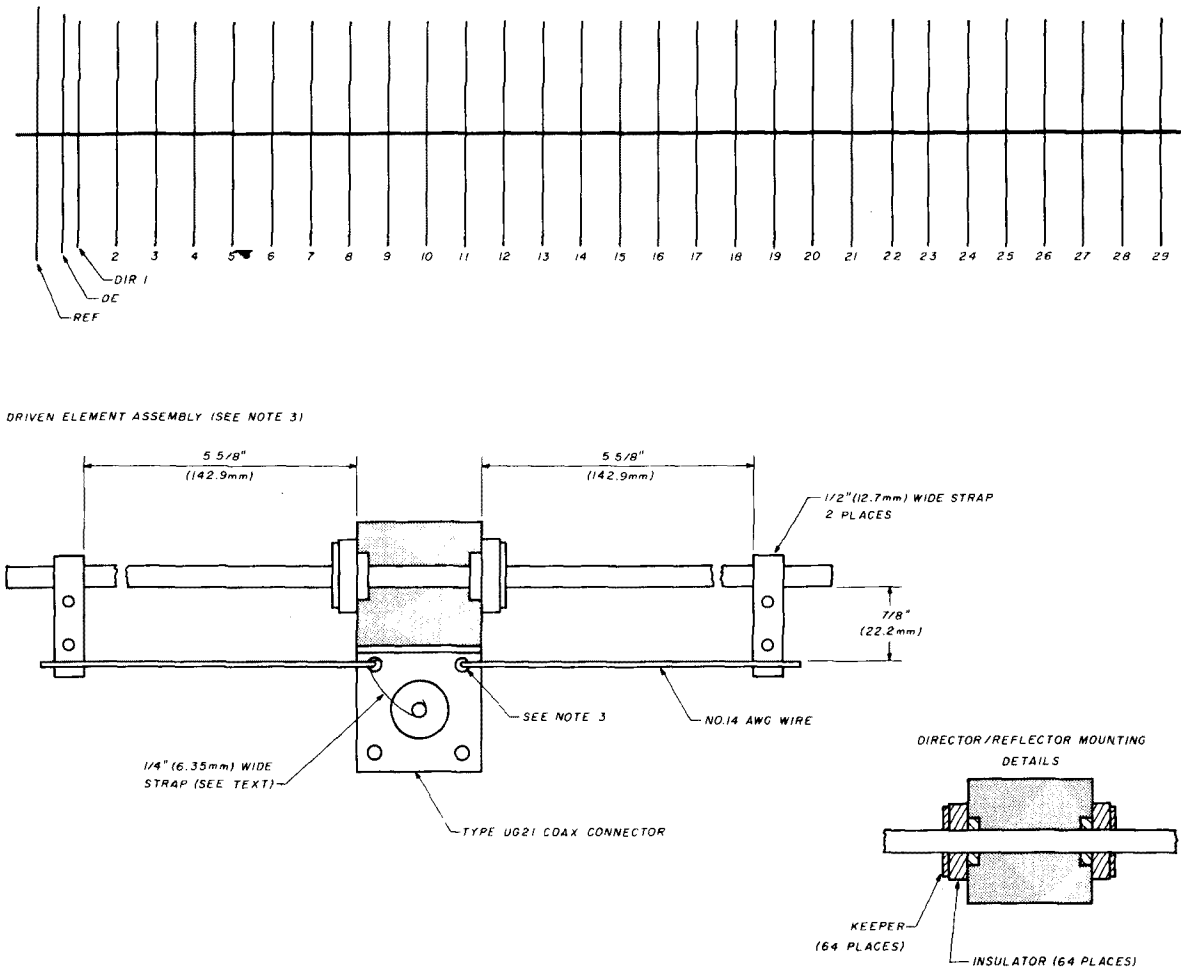


fig. 3. The construction of the high-performance, 31-element, 432-MHz, 24-foot boom long Yagi described in the text and plotted in *fig. 2*. Any deviations from the construction details, dimensions, or the materials specified may decrease performance unless proper compensations are made.

designs, is also used. It's *very* important that a wide (0.25-inch) strap be placed between the center pin on the coax connector and the one side of the balun if the balance of the feed system is to be retained.

All dimensions shown in **fig. 3** *should be copied as closely as possible if performance is to be duplicated. Changing the boom diameter, mounting methods, or element diameter will require the entire design to be modified accordingly and performance can no longer be guaranteed.*

This long Yagi will have about 3 inches of sag if constructed from the materials shown. While lack of boom support will only slightly decrease gain, the mechanical strength will suffer with wind or ice loading. Therefore, I recommend that you support the boom by using either Phyllistran™ or equivalent cable or truss the boom with two pieces of aluminum tubing at least 6 to 8 feet long.

stacking recommendations

This particular design should stack very well. Beamwidth is approximately 20 to 22 degrees in both the E and H planes. The sidelobes are low (17 to 18 dB below the main beam). Therefore the maximum stacking distance (as calculated per reference 13) should be approximately 2.7 wavelengths (74 inches) in the E plane and 2.6 wavelengths (71 inches) in the H plane, but isn't that critical.

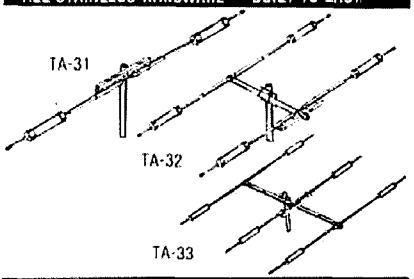
For optimum stacking results the phasing lines should be made from low-loss coaxial cable[12], be as short as possible, and be cut in odd multiples of quarter wavelength[14]. Recommended feeding configurations are described in reference 14. The estimated gain of four of these Yagis properly stacked is 24 to 24.5 dBi, or approximately equivalent to a 16-foot diameter parabolic dish.

summary

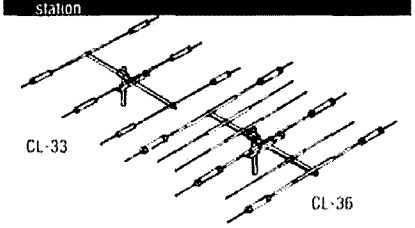
Long boomlength Yagi antennas are now a reality without compromising gain or sidelobe levels. Although bet-

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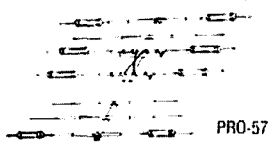
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(part of fig. 3)

Element	Length in inches (Note 1)	Spacing in inches (Notes 1 & 2)
Reflector	13.858	0.000
Driven Element	13.500	5.000
Director 1	12.795	6.913
Director 2	12.362	11.830
Director 3	12.126	17.732
Director 4	12.008	24.617
Director 5	11.890	32.294
Director 6	11.654	40.573
Director 7	11.535	49.234
Director 8	11.496	58.277
Director 9	11.378	67.731
Director 10	11.260	77.567
Director 11	11.221	87.812
Director 12	11.221	98.249
Director 13	11.181	108.877
Director 14	11.102	119.697
Director 15	11.063	130.707
Director 16	11.023	141.718
Director 17	11.023	152.729
Director 18	10.984	163.739
Director 19	10.906	174.750
Director 20	10.748	185.761
Director 21	10.630	196.771
Director 22	10.630	207.782
Director 23	10.630	218.793
Director 24	10.551	229.803
Director 25	10.551	240.814
Director 26	10.551	251.825
Director 27	10.472	262.835
Director 28	10.472	273.846
Director 29	10.472	284.857

Note 1: All lengths are given in inches to 0.001 inch. For best performance, tolerances should be kept to ±0.04 inches maximum, with ±0.02 inches preferred.

Note 2: All spacings are referenced from the reflector to prevent tolerance buildup. It is suggested that you first mark off the reflector position about 1.0 inches from the rear of the boom. Then measure all the spacings from this reference point as shown in the table.

ter designs may be forthcoming, they won't significantly increase the gain or decrease the sidelobe levels of this design *if the boomlength is kept the same.*

Properly designed and constructed Yagi antennas offer very high gain without the windload of a parabolic dish offering comparable gain. In fact, backyard 70-cm EME operation with four of these Yagis is now possible without arousing your neighbor's suspicion!

The antenna described in this month's column is a real performer. Easily duplicated by any interested Amateur, its cost is moderate, especially if local Amateurs join together to buy the materials in quantity. Just think of the satisfaction you'll have in building your own high-gain antenna system!

acknowledgments

This particular Yagi design would not perform as well as it does if it weren't for the help and encouragement of Stan Jaffin, WB3BGU, and Steve Powlishen, K1FO. In particular, I want to thank Steve for optimizing the final design and convincing me that he could in fact predict the performance of a given Yagi design using computer modeling techniques.

'60

new records

At the end of last month's column I alluded to a new 23-cm (1296 MHz) DX record. Since then I have been able to verify that a new record was indeed established on August 13, 1986, when WB6NMT, in San Diego, had a two-way SSB QSO with KH6HME, who was operating from the side of Mauna Loa on the island of Hawaii. This extends the 23-cm DX record worldwide to 2528 miles. Congratulations to Louis and Paul.

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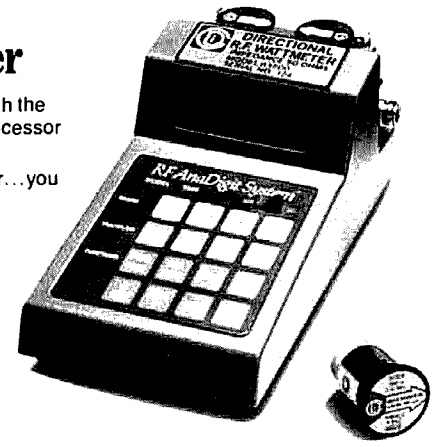
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December 21 (± 1 month)	*Winter peak of sporadic E propagation*
December 22	*Predicted peak of the Ursids meteor shower, 0400 UTC*
December 30	*EME perigee*
January 3	*Predicted peak of the Quadrantids meteor shower, 1830 UTC*
January 10/12	*ARRL VHF Sweepstakes Contest*
January 28	*EME perigee*

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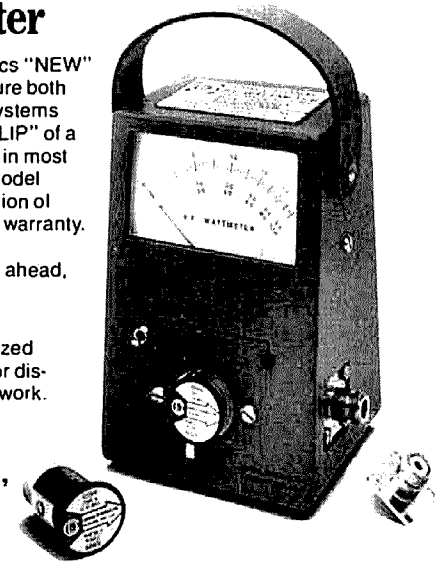


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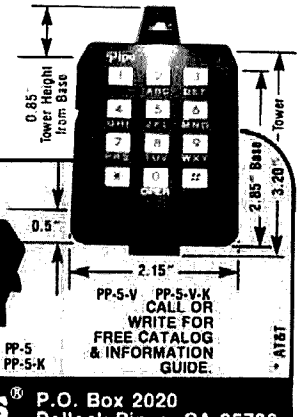
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Ever have a project that made you feel frustrated, yet pleased? Aggravated, but delighted? Unless you've had plenty of experience, all of these emotions will arise the first time you try to rewind a transformer.

Rewinding a transformer by hand is practical if the transformer's output voltages are below 300 to 400 volts. Winding for voltages below 50 is safe and more suitable than winding for voltages above 500, which creates a series of circumstances not addressed by this article. *High-voltage transformers should be wound only by those with previous experience.*

Although rewinding transformers presents an interesting challenge, books and magazines rarely offer sufficient information on the subject. The computer program included here (**fig. 1**), however, should help.

selecting core material

Before any winding project can begin, core material must be selected. This can be a problem, since new core material is usually not available in quantities of less than 85 pounds.* One pound or less is usually sufficient for most homebrewed transformer projects.

Core material is rated according to type, thickness, tongue width, and the number of flux lines it will support. Soft iron (usually silicon steel) is most commonly used for power transformers because of its lower magnetic retentivity. ("Lower retentivity" means that the material retains very little magnetizing orientation after the magnetizing force — i.e., the primary input power — is removed.)

The number of magnetic flux lines per unit area represents the magnetizing force. Flux ratings are typically given in kilogauss (BM) per square centimeter. (If given in inches, the rating would be in kilolines per square inch.) The flux density of silicon steel typically ranges from 12,000 to 20,000 lines per square centimeter. A flux density rating for a used core from a power transformer could safely be placed at about 15,000 kilogauss for calculations.

Core material should be thin to keep eddy currents to a minimum. Eddy currents are currents induced into the metal core material, causing a residual leakage current to flow in the primary winding. A changing magnetic field around the core material, which is a conductor, will induce a voltage into the core, causing a current to flow if there is a closed circuit path. Making the material thinner and insulating its surface increases the resistance to eddy current flow. The use of E and I cut material takes advantage of the higher resistance path in addition to making it easier to install the laminations into the finished coil. Transformer efficiency is increased by reducing the power lost to eddy-current core heating. Power lost in the core is power drawn from the source but not delivered to the load.

Core material is usually measured along the tongue, or center leg or element of an "E." Tongue widths vary from about 1/8 to 4 inches.

The application and amount of transformer power dictate the core size to use. Data specific to a manufacturer's core material must be obtained from the appropriate data sheet. **Table 1** shows a comparison of typical core data as a function of tongue width. The data listed represents M6 silicon steel lamination 0.014-inch thick in an interleaved square stack. The volt-ampere (VA) rating indicates the amount of power the core can handle in a square stack. (In a square stack, core height is equal to tongue width.) Core area (AC) in the table is the numerical square of the tongue width. The remaining values of core loss (CL), copper regulation (CR), and copper loss (RC) were developed in parallel with the VA and tongue width values to support the computer program calculations.

*Ram Sales, 14823 Aetna, Van Nuys, California 91411, sells 29-M6 silicon steel lamination material in one-pound increments. Tongue widths available: 3/8, 1/2, 5/8, 3/4, 7/8, 1, and 1-3/4 inches. Prices range from $1.50 to $3.60 per pound.

By Hugh Wells, W6WTU, 1411 18th Street, Manhattan Beach, California 90266

fig. 1. Power transformer design program. (For enlarged copy of program, send SASE to *ham radio*, Greenville, NH 03048.)

```
10 REM POWER TRANSFORMER DESIGN DATA
20 REM DEVELOPED BY HUGH WELLS DEC 1985
30 BM=14.4:REM FLUX DENSITY IN KILOGAUSS
40 REM
50 GOTO 720
60 REM WIRE TABLE
70 IF CM<=8 THEN WS=41:TI=250:SF=0.85:RETURN
80 IF CM<=10 THEN WS=40:TI=239:SF=0.86:RETURN
90 IF CM<=12 THEN WS=39:TI=215:SF=0.86:RETURN
100 IF CM<=16 THEN WS=38:TI=193:SF=0.87:RETURN
110 IF CM<=20 THEN WS=37:TI=170:SF=0.87:RETURN
120 IF CM<=25 THEN WS=36:TI=155:SF=0.87:RETURN
130 IF CM<=31 THEN WS=35:TI=140:SF=0.88:RETURN
140 IF CM<=40 THEN WS=34:TI=124:SF=0.88:RETURN
150 IF CM<=50 THEN WS=33:TI=110:SF=0.88:RETURN
160 IF CM<=64 THEN WS=32:TI=98:SF=0.88:RETURN
170 IF CM<=80 THEN WS=31:TI=88:SF=0.88:RETURN
180 IF CM<=100 THEN WS=30:TI=80:SF=0.89:RETURN
190 IF CM<=128 THEN WS=29:TI=71:SF=0.89:RETURN
200 IF CM<=160 THEN WS=28:TI=64:SF=0.89:RETURN
210 IF CM<=202 THEN WS=27:TI=57:SF=0.89:RETURN
220 IF CM<=253 THEN WS=26:TI=52:SF=0.89:RETURN
230 IF CM<=320 THEN WS=25:TI=47:SF=0.9:RETURN
240 IF CM<=404 THEN WS=24:TI=42:SF=0.9:RETURN
250 IF CM<=511 THEN WS=23:TI=37:SF=0.9:RETURN
260 IF CM<=640 THEN WS=22:TI=33:SF=0.9:RETURN
270 IF CM<=812 THEN WS=21:TI=30.SF=0.9:RETURN
280 IF CM<=1020 THEN WS=20:TI=2!:SF=0.9:RETURN
290 IF CM<=1290 THEN WS=19:TI=23:SF=0.9:RETURN
300 IF CM<=1620 THEN WS=18:TI=21:SF=0.9:RETURN
310 IF CM<=2050 THEN WS=17:TI=19:SF=0.9:RETURN
320 IF CM<=2580 THEN WS=16:TI=17:SF=0.9:RETURN
330 IF CM<=3260 THEN WS=15:TI=15:SF=0.9:RETURN
340 IF CM<=4110 THEN WS=14:TI=13:SF=0.9:RETURN
350 IF CM<=5178 THEN WS=13:TI=12:SF=0.9:RETURN
360 IF CM<=6530 THEN WS=12:TI=10:SF=0.9:RETURN
370 IF CM<=8234 THEN WS=11:TI=9:SF=0.9:RETURN
380 IF CM<=10380 THEN WS=10:TI=8:SF=0.9:RETURN
390 IF CM<=13100 THEN WS=9:TI=7:SF=0.9:RETURN
400 IF CM>13100 THEN WS=0:RETURN
410 REM CORE RATINGS
420 IF AC<=0.02 THEN AC=0.016:VA=0.02:AW=0.03:CL=0.01:CR=98:RC=220:RETURN
430 IF AC<=0.04 THEN AC=0.035:VA=0.08:AW=0.05:CL=0.02:CR=97:RC=129:RETURN
440 IF AC<=0.07 THEN AC=0.063:VA=0.2:AW=0.12:CL=0.05:CR=96:RC=60:RETURN
450 IF AC<=0.15 THEN AC=0.141:VA=0.9:AW=0.17:CL=0.2:CR=95:RC=50:RETURN
460 IF AC<=0.27 THEN AC=0.25:VA=3:AW=0.19:CL=0.3:CR=75:RC=44:RETURN
470 IF AC<=0.33 THEN AC=0.32:VA=4.5:AW=0.24:CL=0.4:CR=62:RC=39:RETURN
480 IF AC<=0.4 THEN AC=0.39:VA=7:AW=0.29:CL=0.6:CR=52:RC=35:RETURN
490 IF AC<=0.5 THEN AC=0.47:VA=10:AW=0.35:CL=0.8:CR=44:RC=32:RETURN
500 IF AC<=0.6 THEN AC=0.56:VA=14:AW=0.42:CL=1:CR=30:RC=26:RETURN
510 IF AC<=0.8 THEN AC=0.77:VA=25:AW=0.57:CL=1.6:CR=20:RC=22:RETURN
520 IF AC<=1.1 THEN AC=1:VA=40:AW=0.75:CL=2.3:CR=15:RC=19.8:RETURN
530 IF AC<=1.3 THEN AC=1.27:VA=60:AW=0.95:CL=3.4:CR=11.5:RC=15.7:RETURN
540 IF AC<=1.6 THEN AC=1.56:VA=100:AW=1.17:CL=4.7:CR=10:RC=14.2:RETURN
550 IF AC<=2 THEN AC=1.89:VA=150:AW=1.42:CL=6.8:CR=6.5:RC=11:RETURN
560 IF AC<=2.3 THEN AC=2.25:VA=200:AW=1.69:CL=8.1:CR=6:RC=10:RETURN
570 IF AC<=3.1 THEN AC=3.06:VA=300:AW=2.3:CL=11.5:CR=5:RC=9.5:RETURN
580 IF AC<=4.6 THEN AC=4.5:VA=700:AW=3.87:CL=23.1:CR=3.5:RC=6.4:RETURN
590 IF AC<=5.1 THEN AC=5.06:VA=1000:AW=4.4:CL=27.6:CR=3:RC=5.7:RETURN
600 IF AC<=6.3 THEN AC=6.25:VA=1500:AW=4.69:CL=37.2:CR=2.6:RC=5.1:RETURN
610 IF A>7 THEN PRINT "CORE IS TOO LARGE FOR THIS PROGRAM":END
620 REM CALCULATE WINDOW/LATER DATA
630 HA=SQR(AW/3.5):REM CALC WINDOW HEIGHT
640 WB=3.5*HA:REM CALC WINDOW WIDTH
650 HA=INT(HA*1000+0.5)/1000
660 WB=INT(WB*1000+0.5)/1000
670 TL=WB*TI*SF:REM CALC # TURNS PER LAYER
680 TL=INT(TL-0.5)
690 LA=NN/TL:REM CALC # LAYERS
700 LA=INT(LA*10)/10
710 RETURN
720 PRINT CHR$(125):REM CLEAR SCREEN/HOME
730 PRINT :PRINT :PRINT "    POWER TRANSFORMER DESIGN DATA"
740 PRINT :PRINT "ENTER THE VALUES AS REQUESTED---"
750 PRINT :PRINT "CORE TONGUE DIMENSIONS IN INCHES ---"
760 PRINT "   TONGUE WIDTH ";:INPUT W
770 PRINT "          HEIGHT ";:INPUT H
780 PRINT :PRINT "ENTER PRIMARY VOLTAGE ";:INPUT EP
790 PRINT :PRINT "ASSUMING A SINGLE SECONDARY WINDING,"
800 PRINT "ENTER THE DESIRED OUTPUT VOLTAGE";:INPUT ES
805 IF ES>300 THEN PRINT :PRINT "CAUTION: HIGH VOLTAGE IS DANGEROUS":PRINT
"FOR HOME TRANSFORMER WINDING."
810 PRINT :PRINT "ENTER SERVICE RATING # --"
820 PRINT "  1 - INTERMITTANT DUTY (75C RISE)."
830 PRINT "  2 - COMMERCIAL DUTY (50C RISE)."
840 PRINT "  3 - MILITARY DUTY (25C RISE)."
850 PRINT :PRINT "ENTER # ";:INPUT SE
860 IF SE=1 THEN PR=300
870 IF SE=2 THEN PR=500
880 IF SE=3 THEN PR=1000
890 IF SE<1 OR SE>3 THEN 810
900 A=W*W:GOSUB 430
910 REM PERFORM CALCULATIONS
920 NP=(3.49*1000*EP)/(60*AC*0.9*BM):REM CALC PRIMARY TURNS
930 LB=VA/30:LB=INT(LB*100)/100:REM CALCULATES APPROX CORE WEIGHT
940 CP=(4*RC*(NP^2)*(IP^2)):REM CALCULATES COPPER LOSS
950 EF=(VA*100)/(VA+CL+CP):EF=INT(EF*100)/100:REM CALC EFFICIENCY
960 IP=(VA)/((EF/100)*EP):REM CAL PRIMARY CURRENT
970 CM=PR*IP:GOSUB 60:REM CAL WIRE SIZE
980 NS=NP*(ES/EP)*((CR/100)+1):REM CAL SEC TURNS
990 NP=INT(NP*10)/10:NS=INT(NS*10)/10
1000 IP=INT(IP*1000)/1000
1010 TV=NP/EP:TV=INT(TV*10)/10:REM CAL TURNS PER VOLT
1020 IS=VA/ES:IS=INT(IS*1000)/1000:REM CAL SEC CURRENT
1030 NN=NP
1040 GOSUB 620
1050 PRINT CHR$(125):REM CLEAR SCREEN/HOME
1060 PRINT :PRINT "POWER TRANSFORMER DESIGN DATA"
1070 PRINT :PRINT "RATING IS ";
1080 IF SE=1 THEN PRINT "INTERMITTANT DUTY (75C RISE)"
1090 IF SE=2 THEN PRINT "COMMERCIAL DUTY (50C RISE)"
1100 IF SE=3 THEN PRINT "MILITARY DUTY (25C RISE)"
1110 PRINT :PRINT "PRIMARY VOLTAGE = ";EP;" VOLTS."
1120 PRINT "PRIMARY CURRENT = ";IP;" AMPS."
1130 PRINT "# PRIMARY TURNS = ";NP
1140 PRINT "PRI TURNS PER VOLT= ";TV;" : 1"
1150 PRINT "PRIMARY WIRE SIZE = #";WS
1160 PRINT LA;" LAYERS OF ";TL;" TURNS/LAYER."
1170 PRINT :PH=LA/TI:REM CALCULATES HEIGHT OF PRIMARY
1180 PRINT "SECONDARY VOLTAGE = ";ES;" VOLTS."
1190 PRINT "MAX SEC CURRENT = ";IS;" AMPS."
1200 PRINT "# SECONDARY TURNS = ";NS
1210 NN=NS
1220 CM=PR*(VA/ES):REM CALC SECONDARY WIRE SIZE
1230 GOSUB 60:REM CALC SECONDARY WIRE SIZE
1240 GOSUB 620:REM CALCULATE TURNS/LAYER
1250 PRINT "SECONDARY WIRE SIZE = #";WS
1260 PRINT LA;" LAYERS OF ";TL;" TURNS/LAYER."
1270 SH=LA/TI:REM CALCULATES HEIGHT OF SECONDARY
1280 TH=PH+SH:REM CALCULATES TOTAL HEIGHT OF WINDING
1290 IF TH>=(HA*0.8) THEN PRINT :PRINT "WARNING: WINDOW HEIGHT EXCEEDED--"
1300 REM TH=TOTAL HEIGHT - 20% OF SPACE ALLOWED FOR INSULATION
1310 PRINT :PRINT "ENTER 1 TO CONTINUE RATINGS";:INPUT Z
1320 PRINT CHR$(125):REM CLEAR SCREEN/HOME
1330 PRINT :PRINT "GENERAL TRANSFORMER RATINGS"
1340 PRINT :PRINT "APPROX POWER RATING = ";VA;" VA."
1350 PRINT "EFFICIENCY      = ";EF;"%"
1360 PRINT "COPPER REGULATION = ";CR;"%"
1370 PRINT "APPROX CORE LOSS = ";CL;" WATTS."
1380 PRINT "APPROX CORE WEIGHT = ";LB;" LBS."
1390 PRINT "ASSUMED FLUX DENSITY= ";BM;" KILOGAUSS"
1400 PRINT "SOURCE FREQUENCY = 60 HERTZ."
1410 PRINT "ASSUMED CORE WINDOW RATIO = 3.5:1"
1420 PRINT "WINDOW AREA = ";AW;" SQ INCHES."
1430 PRINT "WINDOW WIDTH = ";WB;" INCHES."
1440 PRINT "WINDOW HEIGHT = ";HA;" INCHES."
1450 PRINT :PRINT "DO ANOTHER? 1=Y 2=N ";:INPUT Z
1460 IF Z<>2 THEN RUN
1470 END
```

The window area (i.e., space allowed for the coil) of a core, another important measurement, usually follows the tongue width in size. Although window width and height may vary from one manufacturer to another, window area appears to remain fairly uniform for a given core size.

The program assumes the typical width-to-height ratio of 3.5:1. For instance, if the winding width is 0.875 inch and the window area is 0.219 square inches, the window height will be 0.25 inch. The number of turns-per-layer and the number of layers is determined from the window width and area. For example, the number of turns-per-layer is found by multiplying window width by the number of turns-per-inch and by the winding space factor. Then the number of layers is found by dividing the total number of turns-per-winding by the number of turns-per-layer. Ultimately, the number of turns of wire that can be wound onto the core will be determined by wasted space, wire bulging, and insulation thickness. This may sound like a lot, but one must keep in mind that only so much window area is available to accommodate the wire.

salvaging a used core

Though it's possible to start from scratch and build your own transformer, it's more fun to tear down an old transformer and rewind the old core. Recovering an old core, however, can be tricky. Perhaps the slowest, most time-consuming chore is cleaning out the old potting compound. Most transformers have been vacuum-impregnated with material that affords protection from moisture and keeps laminations from "singing" by gluing them together. If it's one of the tougher materials such as Hysol or one of the new urethanes, it's not easily removed by anything short of a jackhammer. Such transformers are best set aside — perhaps for use as door stops.

Transformers potted in wax are the easiest to salvage. Wax and some varnishes can be removed by soaking the transformer in paint or varnish remover, and/or a suitable solvent. It's necessary to soften the potting compound sufficiently to allow the laminations to be removed; vacuum impregnation is very efficient and lengthy soaking may be required before the laminations will be ready to slide apart, even with the ad-

Table 1. Core table for silicon steel.

Tongue Width (Inches)	Sq stack Core area (AC)	Volt-ampere Rating (VA)	Sq inch Window area (AW)	Core Loss (CL)	Copper Regulation (CR)	Copper Loss Factor (RC)
0.126	0.016	0.02	0.03	0.01	98	220
0.187	0.035	0.08	0.05	0.02	97	129
0.25	0.063	0.2	0.12	0.05	96	60
0.375	0.141	0.9	0.17	0.2	95	50
0.5	0.25	3.0	0.19	0.3	75	44
0.566	0.32	4.5	0.24	0.4	62	39
0.625	0.39	7.0	0.29	0.6	52	35
0.686	0.47	10.0	0.35	0.8	44	32
0.75	0.56	14.0	0.42	1.0	30	26
0.875	0.77	25.0	0.57	1.6	20	22
1.0	1.0	40.0	0.75	2.3	15	19.8
1.125	1.27	60.0	0.95	3.4	11.5	15.7
1.25	1.56	100.0	1.17	4.7	10	14.2
1.375	1.89	150.0	1.42	6.8	6.5	11
1.5	2.25	200.0	1.69	8.1	6.0	10
1.75	3.06	350.0	2.3	15.9	5.0	9.5
2.125	4.5	700.0	3.87	23.1	3.5	6.4
2.25	5.06	1000.0	4.4	27.6	3.0	5.7
2.5	6.25	1500.0	4.69	37.2	2.6	5.1

ded encouragement of a hammer. Work outdoors — and have patience. It takes time and plenty of elbow grease to remove the debris.

If none of the windings are to be retained, you can partially remove some of the wire by cutting through the coil with a hacksaw before soaking. If you do this, be sure to protect the core from mechanical damage. Be careful, too, with the paper tube or winding bobbin; it will be used again in rewinding the core.

When salvaging the core, it's usually best to remove all of the wire. However, there are circumstances in which you might want to retain one or more of the windings. In these cases, the salvage operation is more difficult because the transformer can't be soaked in paint remover or solvent. If some of the windings are to be saved, the tedious chore of removing the laminations without damaging the wire becomes necessary.

In this procedure, a wood splinter is usually driven between the paper core form and the core to prevent coil vibration. Remove the splinter to free up space to work in, then separate and remove the outer laminations from the others with a thin knife to provide additional room. If the outer laminations are damaged during removal, correct or remove small dents in the metal by tapping lightly with a hammer against a flat surface. (Because silicon steel is almost as soft as most modern automobile fenders, it's sometimes referred to as "fender iron," and the process affectionately titled "body and fender work.") The loss of one or two laminations won't affect the operation of the rewound transformer significantly; any significant loss, however, will reduce the power rating (VA) of the core. You can compensate for this by determining the remaining stack height for entry into the computer's calculation.

The wire sizes used on the old transformer provide an important clue as to the amount of power the original transformer was able to deliver. The power can be estimated by multiplying the secondary winding current by the winding's voltage. For transformers with multiple secondary windings, add up the individual winding power to get the total transformer power. To determine the nominal current value a wire can handle, assume 500 CM per ampere. For example, a No. 23 wire will handle 1 ampere of current. Wire has a resistance value which contributes to heat loss as current flows through it. For a given current value, a larger wire size would dissipate less heat because of its lower resistance.

wire sizes

Table 2 provides comparative data for various wire sizes. Of concern is the amount of current a particular wire will carry and how much space is required when the wire is used for windings on a transformer. Heat rise is affected by wire size and is controlled by the circular mil area (CM) of the wire. For example, a wire of 1000 CM carrying 1 ampere of current will run cool with about a 25-degree C temperature rise

Table 2. Wire table.

Circular mil area (CM)	Wire size (WS)	Turns/inch (TI)	Space factor (SF)
8	41	250	0.85
10	40	239	0.86
12	39	215	0.86
16	38	193	0.87
20	37	170	0.87
25	36	155	0.87
31	35	140	0.88
40	34	124	0.88
50	33	110	0.88
64	32	98	0.88
80	31	88	0.88
100	30	80	0.89
128	29	71	0.89
160	28	64	0.89
202	27	57	0.89
253	26	52	0.89
320	25	47	0.90
404	24	42	0.90
511	23	37	0.90
640	22	33	0.90
812	21	30	0.90
1020	20	26	0.90
1290	19	23	0.90
1620	18	21	0.90
2050	17	19	0.90
2580	16	17	0.90
3260	15	15	0.90
4110	14	13	0.90
5178	13	12	0.90
6530	12	10	0.90
8234	11	9	0.90
10380	10	8	0.90
13100	9	7	0.90

above ambient. At 1 ampere, 500 CM will run warmer with a 50-degree C rise, and 300 CM will run hot at a 75-degree C rise. Enamel insulation (purple to black color) on copper wire must be operated below 50 degrees C rise to keep the insulation from disintegrating. Formvar (orange color) and the newer high-temperature insulation materials are capable of handling the higher temperature rises without failure.

The number of turns-per-inch (TI) and space factor (SF) are provided in **table 2**. The TI number is based upon the diameter of double-coated formvar insulation. Wire coated with enamel will also be of about the same diameter. The SF is a judgment of wasted space at the end of each layer and is dependent upon winding uniformity. Working together, TI and SF values improve the chances of predicting whether or not all of the desired number of turns of the selected wire will fit into the core window. It's better to know the fit up front than after the core has been wound.

core stacking

There are two ways of stacking *E* and *I* material to make up a transformer core. The first way is an interleaved stack, which means that *E* laminations are alternated in direction as they're stacked to create a double closed loop (i.e., with no air gap at the core ends). Individual *I* laminations are pressed into the gaps between the *E's* to fill in the core. The second way is called a butt stack where all of the *E's* are stacked in parallel to maintain the *E* shape. The ends of the *E's* are butted up against a stack of *I* laminations equally as high to form a nearly closed core.

Each stacking method is best suited for a particular application. An interleaved stack is most suitable when the transformer windings support only alternating current power. With only alternating current power applied, a transformer's efficiency is improved with a closed core which keeps the flux lines contained within the core. However, the tendency for core saturation increases with the closed core, particularly with the addition of direct current in one of the windings. When used with a rectifier, the non-uniform waveform of the pulsating dc produced will affect the transformer's performance, but the effect is usually ignored.

If it's necessary to have direct current flowing in one of the windings, a butt stack is preferred over an interleaved stack. Direct currents cause a biasing of the magnetic flux in the core, and if too high, the core could saturate. That means the number of flux lines has been raised to the maximum density the core can handle. (A simple analogy: it's like trying to put more people on a bus than the bus can hold. After the bus is filled, adding another person to the bus causes someone to be squeezed out elsewhere. The bus is "saturated.") When saturated, the core will fail to respond to the AC changes applied to the primary winding. Therefore, the advantage of the butt stack is that a small air gap can be created between the ends of the *E* and *I* stacks to provide a flux leakage path. Flux leakage reduces the tendency for core saturation by allowing some of the flux lines to escape. A thin piece of paper or tape can be used to establish and hold a desired gap dimension.

winding a core

Winding wire onto a core may be accomplished by either hand or machine. Most home experimenters use the hand-winding technique. Whether hand- or machine-wound, the wire insulation must be treated with care to prevent damage; scratches in the insulation can cause the turns of wire to short. Because of potential insulation damage, wire salvaged from used transformers is seldom used for new transformers.

When starting a winding operation, wind the wire on a bobbin or paper tube. The bobbin or paper tube serves as a smooth winding surface for the wire and provides insulation from the core. The form also provides support for the coil before the core laminations

```
     POWER TRANSFORMER DESIGN DATA

ENTER THE VALUES AS REQUESTED---

CORE TONGUE DIMENSIONS IN INCHES ---
   TONGUE WIDTH
             0.875
         HEIGHT
             0.875

ENTER PRIMARY VOLTAGE
                    120

ASSUMING A SINGLE SECONDARY WINDING,
ENTER THE DESIRED OUTPUT VOLTAGE
                                    12
ENTER SERVICE RATING # --
  1 - INTERMITTANT DUTY (75C RISE).
  2 - COMMERCIAL DUTY (50C RISE).
  3 - MILITARY DUTY (25C RISE).

ENTER #
       2

POWER TRANSFORMER DESIGN DATA

RATING IS
COMMERCIAL DUTY (50C RISE)

PRIMARY VOLTAGE = 120 VOLTS.
PRIMARY CURRENT = 0.221 AMPS.
# PRIMARY TURNS = 699.4
PRI TURNS PER VOLT= 5.8 : 1
PRIMARY WIRE SIZE = #29
7.9 LAYERS OF 88 TURNS/LAYER.

SECONDARY VOLTAGE = 12 VOLTS.
MAX SEC CURRENT = 2.083 AMPS.
# SECONDARY TURNS = 83.9
SECONDARY WIRE SIZE = #19
2.9 LAYERS OF 28 TURNS/LAYER.

ENTER 1 TO CONTINUE RATINGS

GENERAL TRANSFORMER RATINGS

APPROX POWER RATING = 25 VA.
EFFICIENCY      = 93.98%
COPPER REGULATION = 20%
APPROX CORE LOSS = 1.6 WATTS.
APPROX CORE WEIGHT = 0.83 LBS.
ASSUMED FLUX DENSITY= 14.4 KILOGAUSS
SOURCE FREQUENCY = 60 HERTZ.
ASSUMED CORE WINDOW RATIO = 3.5:1
WINDOW AREA = 0.57 SQ INCHES.
WINDOW WIDTH = 1.412 INCHES.
WINDOW HEIGHT = 0.404 INCHES.

DO ANOTHER? 1=Y 2=N
```

fig. 3. Sample screen printout of design.

sulation cracking could occur, exposing the copper and increasing the possibility of shorts.

The coil is finished off with a layer of paper or tape to provide a protective cover for the wire. Then the laminations are stacked, using either the interleaved or butt stack method. When using the interleaved method, alternate the direction of the *E's* as they're inserted into the tube or bobbin, then slip the *I's* into the open slots. If using the butt stack, slide all of the *E's* into the form from one side. Stack the *I* laminations to an equal height of the *E's* and hold the *I's* together either with a small amount of wax or varnish. A narrow strip of masking tape may be wrapped around the *I* stack. The tape must be positioned so that it won't interfere with ends of the *E* stack. Tape may be wrapped around the cores edge (not over the coil) to hold the assembly together. Apply power and check out the completed transformer. Unloaded, the output voltage will be slightly higher than the calculated value. It will, however, decrease under load.

the program

The program (**fig. 1**) will calculate the number of turns and wire size for a given core. Both are based upon the maximum power the core can handle (assuming a 90-percent power factor). To keep the winding project simple but usable, and within the grasp of a novice winder, the program was written to accommodate a single secondary winding. For multiple windings, the program is used to find the amount of power (VA) available that must be distributed among all windings. Likewise, the wire size must decrease so that the copper mass will fit within the available core window area.

Formulas used in the program are based on a floating point decimal. Some program adjustment may be required for machines operating with INTEGER BASIC.

The program was written in straightforward BASIC so *that it* would be transportable to the greatest number of BASIC dialects. All program variables (in **fig. 2**) have been listed to provide a troubleshooting aid in case of a syntax error. Variables have been limited to two characters to maintain program simplicity and to accommodate machines limited to two alpha variables. Should that still be a problem, the second character of each variable may be changed to a number. An abundance of REM statements have been used to aid in following each program step.

Program flow is organized with the GOSUB tables at the beginning of the program. In BASIC, the computer always starts at the top of the program (lowest line number) and searches down for a particular GOSUB line. Execution time is reduced when the GOSUB line is found at the beginning of the search. Upon completion of the GOSUB operation, a stack pointer returns the program to the line number following the GOSUB instruction without having to search further. A sample run is shown in **fig. 3**.

Whether you're an educator or an experimenter, this computer program should become a valuable tool. It will provide comparative winding data for various core sizes and winding voltages. Give it a try!

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ham radio

are inserted. Machine-winding on a plastic bobbin, rather than a paper tube, can allow up to a 30-percent increase in the amount of copper that can fit on the core. Hand-winding onto a bobbin is also desirable. However, in the absence of a bobbin, one must use a paper tube. Tubes made commercially are constructed of pressed paper and impregnated with wax or varnish. The coil form (tube or bobbin) may be reused if salvaged along with the core, but burrs and sharp points should be removed from the form with a smooth file or sharp knife to prevent wire from snagging during the winding operation.

If you didn't salvage the paper tube, you can make one from shoe box cardboard. Tubes for small transformers can be made from a manila folder or post card stock. To shape the tube, it's first necessary to stack the core to form an *E* to the desired height, which is usually square (i.e., height is equal to tongue width). Masking tape wrapped around the outer legs will hold the laminations in place while the tube is being formed. Cut a piece of the cardboard long enough to make one turn around the center leg and as wide as the tongue is long. Care must be taken to prevent creases in the cardboard except at the corners of the core. Creases tend to decrease the strength of the tube, causing it to collapse during the wire winding operation. The seam in the tube may be at one of the corners. Knead the cardboard against the iron core to remove any tendency to bulge; you can use the round shaft of a screwdriver as a kneading tool. After the tube is formed and trimmed to fit tightly around the core leg, lay a single strip of masking tape along the seam to cover the gap. With the tube still on the core, wrap one turn of tape around each end of the tube to increase the end strength slightly and to keep the gap closed. Remove the tube from the core and mark the tube sides that were against the flat sides of the lamination. Because the core stack is not perfectly square, marking of the tube is necessary for core re-stacking orientation. After forming, the tube may be soaked with varnish to improve its mechanical stability and moisture resistance.

During winding, it will be necessary to start and stop wire winding on the flat sides of the lamination. It's usually desirable to place the primary winding (the winding with the smallest wire size) next to the core. Because of the shorter distance around the core, less wire is used. Less wire means lower resistance, lower copper loss, and better regulation. Starting and ending coil ends should be on the same side of the core for convenience. Place the wire near one end of the tube with the wire end extending outward about 6 inches. Use a small tab of masking tape like a flag to hold the starting end in place as close to the tube edge as possible. Wind the wire onto the tube one turn at a time over the flag, positioning the wire tightly against the adjacent turns of wire. To avoid wasting space, be careful not to overlap one wire over another.

A	*Core selector*
AC	*Core cross sectional area*
AW	*Window area*
BM	*Flux density*
CL	*Core loss*
CM	*Circular mil area of wire*
CP	*Copper loss*
CR	*Copper regulation*
EF	*Transformer efficiency*
EP	*Primary voltage*
ES	*Secondary voltage*
H	*Stack height*
HA	*Window height*
IP	*Primary current*
IS	*Secondary current*
LA	*Layers of wire*
LB	*Core weight*
NN	*Temporary variable*
NP	*Number of primary turns*
NS	*Number of secondary turns*
PH	*Height of primary winding*
PR	*Temperature rise rating*
RC	*Equiv. series resistance of copper around core*
SE	*Temperature rise selector*
SF	*Layer space factor*
SH	*Height of secondary winding*
TH	*Total winding height*
TI	*Number of turns per inch of wire*
TL	*Number of turns per layer*
TV	*Turns per volt-primary winding*
VA	*Volt-amp rating*
W	*Tongue width (inches)*
WB	*Window width*
WS	*Wire size*
Z	*Temporary variable*

fig. 2. List of program variables.

One of the biggest problems with hand-winding is holding the first and last turn of each layer in place on a paper tube; the end turns tend to fall off because of weak end support. A dab of glue from a glue stick, or of paraffin or rubber cement, on the end two or three turns will help hold them in place.

Each successive layer of wire will tend to fall between the wires of the layer below unless a paper separator between windings is used as a winding surface. Even though it has a slick surface, a strip of wax paper usually works well as a separator. It can be wrapped over the winding layer and tacked in place with a hot soldering iron. Heat from the soldering iron will melt the wax, sticking the paper ends together. The next layer of wire over the separator will hold it permanently in place. Each layer wound onto a paper tube will be progressively shorter (i.e., have fewer turns) than the layer below it, causing the coil to be tapered like the base of a pyramid. Using a bobbin solves the tapering problem.

Keep the wire tight during winding to reduce wasted space and wire bulging. As each turn of wire is added, it's pressed against the form to remove the bulge. In addition, the wire mustn't be stretched because in-



using spreadsheet programs for Amateur Radio projects

Most of us who own home computers have learned to program them in BASIC and found ways to use them in the ham shack. Although BASIC is easy enough to learn and to use, there are alternatives that make some chores much easier. For example, I've been using Lotus 1-2-3™, a highly respected and widely used electronic spreadsheet, for a number of tasks such as keeping and duplicating contest logs, printing QSL cards, and the like. Any of the other popular spreadsheets, such as Multiplan™, would probably work just as well.

Spreadsheet programs are personal productivity tools that allow us to analyze data and create reports. The spreadsheets themselves are large grids of entries that can include text, formulas, and numerics. Since worksheets are able to store formats, formulas, and data, changes can be made to files in order to update or revise data.

I use Lotus 1-2-3 to generate DXCC and WAS records and schedules (**fig. 1**, personalized QSL cards (**fig. 2**), mailing labels, and graphs (**fig. 3**). The QSL application includes a "macro," a list of commands that can be executed as needed in order to make the job of entering information and printing the results largely automatic.

No matter what spreadsheet program you use, you should have no difficulty duplicating the worksheets. The 5BDXCC record (**fig. 1**) is made by entering the names of all countries worked in the first column and placing the various band headings at the top of columns B through F. Using the **@COUNT** function* produces an automatic update of the band totals at the bottom of each column. Subtotals for each country, shown on the right, also use the same **@COUNT** function. As new countries are worked, all that's required is the insertion of new lines where needed to accommodate the new listings. When the **WORKSHEET-INSERT-ROW** command is used, the formulas are revised automatically to count the new totals. The 5BWAS worksheet is basically the same, except that I entered column dividers between each band to improve the appearance of the graph. As with the DXCC record, the program's **@COUNT** or **@SUM** functions have been used to determine the totals.

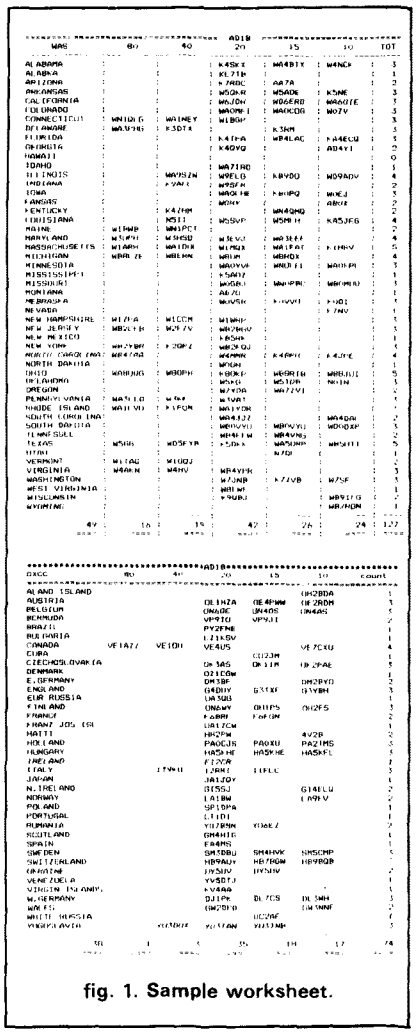


fig. 1. Sample worksheet.

QSL card generator

The QSL card generator is a macro-driven file that allows quick, easy preparation of a large or small number of QSLs. **Figure 2** shows a typical card, printed on inexpensive continuous-feed postcards available at most computer supply stores.

The QSL worksheet is designed in columns A through F, using rows 1 through 18. The print range covers rows 1 through 21, which allows the print head to return to the starting point on the *next* QSL card. All entries are made by means of the data-entry and printing macro. When started, the macro moves the cursor around the card as the return key is struck. This feature is a big time-saver, particularly because the macro sends the card to the printer when the *last entry* is made. The line printer must be turned on when the QSL generator is running because the finished card is sent to the printer after the last of the data is entered.

The construction of the QSL template is fairly simple. Starting with a blank default worksheet, save a file under the name "QSL." Referring to **fig. 2**, use rows 1 through 7 to design the call sign. (Although I used numbers and letters to design my card, any character or keyboard symbol may be used.) Next, design the station report in rows 9 through 11. Insert the message, address, or other information in rows 13 through 18.

When the design is complete, use the TAB key to move to a blank cell of the worksheet and enter the macro. The macro listing shows all of the com-

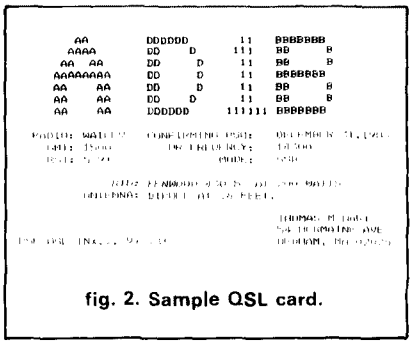


fig. 2. Sample QSL card.

*Commands shown are specific to Lotus 1-2-3. – Ed.

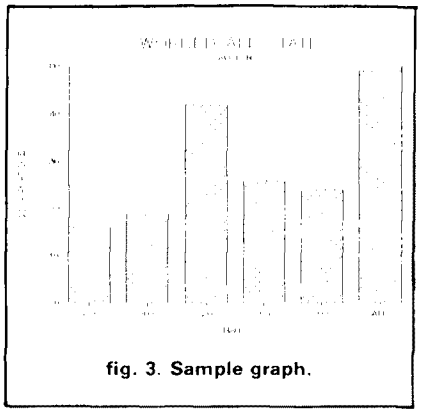


fig. 3. Sample graph.

/
RANGE
NAME
CREATE
"/Z"
"RETURN"
"RETURN"

These commands establish the macro name, allowing us to run the system by typing only two keys, "ALT-Z." Next, enter the macro commands listed in **table 1** and save the completed file. To run a batch of QSL cards, load the QSL file, placing the cursor on the cell where the call sign is to be entered. Start the macro with "ALT-Z" and enter all of the station report data. After each item, hit the "RETURN" key to move to the next. Load the line printer with postcards and turn it on; printing will begin as soon as the last entry is made.

To improve the appearance of the QSLs, I've added a setup string to the macro to emphasize print on the call sign. The two lines that turn the emphasized type on my Epson printer on and off are, respectively, "/027/069" and "/027/070". Other printers will require different instructions; consult your manual.

I've found the QSL generator helpful because it processes data quickly and easily. New designs can be prepared to cover special events, portable operations, and holidays.

label-maker

A simple application, but one that's useful for preparing mailings to club members or the like, is a label-maker. Although routines written in BASIC are often used to generate labels, using Lotus 1-2-3 is simpler, quicker, and more convenient.

I've been using labels consisting of nine rows from top to bottom. This allows plenty of room for even the most complex mailing address. The starting point is the creation of a list of names and addresses in a single column of a worksheet. It's not necessary to set the column width to any particular size because the text will simply run into the next column.

Each new record starts nine rows below the prior starting point. If you select blank mailing labels with only six rows, be sure to change the spacing accordingly. Also remember that any time you enter a street address — "1 Main Street," for example — the program requires that an apostrophe precede the entry to tell the program that text, rather than numerals, is being entered. To create a new file, simply type the names and addresses in a column as you want them to appear on the labels.

To prepare the file for printing, type the following commands:

/
PRINT
PRINT
OPTIONS
OTHER
UNFORMATTED
MARGIN
TOP 0
MARGIN
BOTTOM 0
QUIT
RANGE (SET YOUR RANGE)
QUIT

Next, save the label file under some descriptive name. When you're ready to print, load the label file and enter the following commands:

/
PRINT
PRINT
ALIGN
GO[Note:labels will be generated at this point.]
QUIT

No provision has been made to allow alphabetical sorting. New names can be added to the end of the list; names can be deleted with the program's "WORKSHEET-DELETE-ROW" feature. To print a batch of labels, all that's required is the six keystrokes listed above. Hint: starting printing with the print head set two rows down from the

Table 1. QSL card macro. Note: explanations in right-hand column are not necessary for operation.

```
\Z      {?}~                    ENTER THE CALL SIGN
        {RIGHT}{RIGHT}          MOVE TO THE RIGHT
        {RIGHT}                 MOVE TO THE RIGHT
        {?}~{DOWN}              ENTER THE DATE AND MOVE DOWN
        {LEFT}{LEFT}            MOVE LEFT
        {LEFT}                  MOVE LEFT
        '{?}~                   ENTER THE TIME
        {RIGHT}{RIGHT}          MOVE TO THE RIGHT
        {RIGHT}                 MOVE TO THE RIGHT
        '{?}~{DOWN}             ENTER THE FREQUENCY AND MOVE DOWN
        {LEFT}{LEFT}            MOVE LEFT
        {LEFT}                  MOVE LEFT
        '{?}~                   ENTER THE REPORT
        {RIGHT}{RIGHT}          MOVE RIGHT
        {RIGHT}                 MOVE RIGHT
        {?}~                    ENTER THE MODE
        {LEFT}{LEFT}            MOVE LEFT
        {LEFT}                  MOVE LEFT
        /PPCAARA1.G8~           SEND THE QSL TO THE PRINTER
        OOUMTO~MBO~S            USE CONTINUOUS FEED FORMAT
        \027\069~QG             SEND DOUBLE-STRIKE CHARACTER
        CARA9.G21~              PRINT BALANCE OF QSL
        OOUMTO~MBO~S            USE CONTINUOUS FEED FORMAT
        \027\070~QGQ            CANCEL DOUBLE-STRIKE
        {UP}{UP}                MOVE TO START
        /XG\Z~                  RESTART THE MACRO
```

the ham notebook

mands required to run the application. To start, enter the following in the selected cell:
top of the label will help center the text on the label.

graphs

You can use Lotus 1-2-3 to generate line graphs, bar graphs, stacked bar graphs, pie charts, and such. A graph such as shown in **fig. 3**, requires the following data entry:

80	**16**
40	**19**
20	**42**
15	**26**
10	**24**
ALL	**49**

These data are entered in columns A and B, starting in the first row. The following commands are then entered in the worksheet before saving the final product:

```
/
G
TYPE BAR
X-AXIS A1..A6
A RANGE B1..B6
OPTIONS TITLES
FIRST "WORKED ALL STATES"
TITLES SECOND "AD1B"
TITLES X-AXIS "BAND"
TITLES X-AXIS
  "CONFIRMATIONS"
QUIT
VIEW    [This allows the graph to be seen.]
QUIT
```

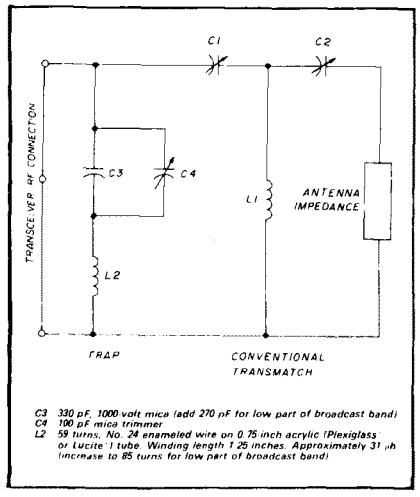
The **VIEW** command allows you to see the graph and print it. You can also save the data while in the GRAPH menu and make a print by using the PRINT GRAPH disk in the Lotus package. **Figure 3** shows the final appearance of the sample graph for my 5BWAS certificate.

Thomas M. Hart, AD1B

suppression of high-level AM broadcast signals

A transmatch and/or receiver input high-pass filter will provide effective attenuation of AM broadcast signals particularly troublesome on the 160-meter band. Transceivers with diode T/R switching, however, are vulnerable to inter-modulation and cross-modulation interference caused by rectification of very high-level signals from close-in, high-power broadcast stations. A transmatch may not provide sufficient attenuation of the offending signal, and filtering in the receiver input after the diodes will not alleviate the problem. A conventional receiver high-pass filter cannot be put ahead of the diode switch because it would be subjected to the 100-watt output of the transceiver.

A method of suppressing a high-level AM broadcast signal is shown in **fig. 1**. C3 and C4 in series with L2 form a series shunt trap tuned exactly to the

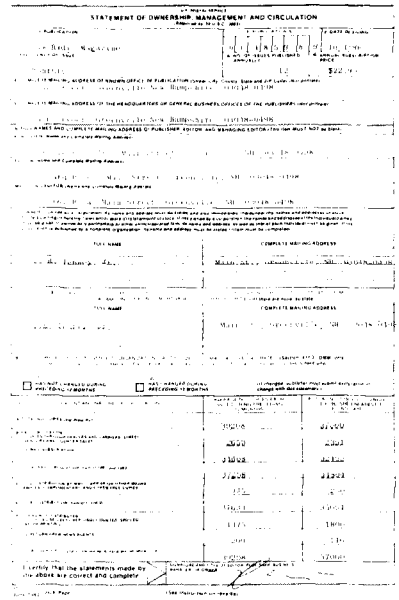


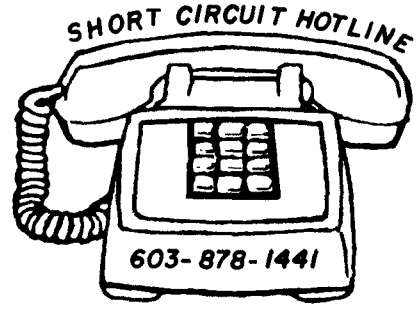
frequency of the broadcast station. L2 is a high-Q coil providing a very low shunt resistance at resonance, thus providing deep attenuation of the broadcast station signal. With the LC trap shown the impedance is high enough above 1.8 MHz so as not to attenuate signals on any of the bands. The high trap impedance also limits the current through L2 on transmit to an insignificant value above 1.8 MHz. In this arrangement the combined trap and transmatch becomes the impedance-matching network. It is necessary only to readjust C1 and C2 for minimum SWR in the normal manner.

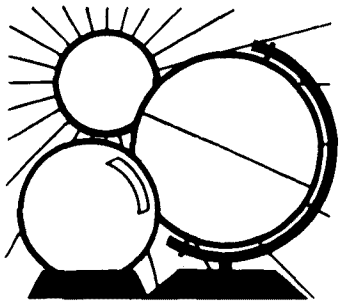
In my case there is a 50-kW station about 2/3 of a mile away on 1500 kHz. The trap reduces the 1500-kHz signal from 2.5 to 0.08 volts. It has cleaned up the 160-meter band dramatically and does not affect received signals above 1.8 MHz. Proper impedance match to the transceiver is achieved on all bands by tuning for minimum SWR in the normal manner. The trap coil is not even warm to the touch after transmitting 100 watts, key down.

Jack Geist, N3BEK

ham radio







DX FORECASTER

Garth Stonehocker, KØRYW

more on winter anomalous absorption

December 1984's DX Forecaster discussed the wintertime increased absorption anomaly that results in periods of weak signals, each lasting a few days. Recent research has produced a more detailed explanation of its cause and effects, as well as a method of prediction.

The cause is not simple, but basically involves the arrival of particles into the polar regions during a geomagnetic disturbance. Their arrival triggers a series of events: the particles ionize nitric oxide (NO) and molecular oxygen (O_2) in the auroral zone's (60- to 70-degree latitude) D and E regions, causing auroral absorption there and electron depletion in these regions for about three days of disturbance.

About two to four days later, the heat buildup from the process produces stratospheric winds that blow the neutral NO and heat to the west (in the northern hemisphere) and to lower latitudes (e.g., 40 degrees). As a result, the NO ionizes in the D region signal absorption areas (1000 to 2000 km in diameter) and spreading, affecting mid-latitude east-west DX propagation paths for the following five to six days.

The absorption develops in two areas on opposite sides of the earth. The areas in between experience lower than normal *winter absorption*. This results in the earth's division into four alternating longitudinal bands: one quarter of the earth's longitudes experiences above-average signal strengths; a second quarter, signal strengths below average; the third, above-average signal strengths, and the fourth, signal strengths below average. This pattern moves about 30 degrees — the equivalent of two time zones — per day. This same pattern appears five to six days later in the lower latitudes (e.g., 40 degrees). However, it does not coincide with the previously described trough condition. The amount of absorption is generally 20 dB, but can be as high as 40 dB for several hours during a day.

From the above discussion, a procedure for forecasting becomes apparent. Geophysical scientists call the warm stratosphere accompanying the anomalous absorption "Stratwarm." Station WWV announces the presence (and location) of Stratwarm conditions at 18 minutes after the hour. During December and January, if a high-value A index is announced, listen for the Stratwarm location. The complementary Stratwarm location will be on the other side of the earth, with lower absorbing regions (depleted NO) in between.

The higher signal strengths in the low-absorption regions are a boon to DXers. A globe or polar map is useful in tracking these regions. Remember, these regions move *west* at a rate of 30 degrees (longitudinally) per day.

Anomalous absorption usually occurs in the months of December and January, with the most useful activity occurring after January 15.

last-minute forecast

The higher frequency bands are expected to be best during the third and fourth weeks of the month, when the probability of a solar flux increase is highest. The resulting increase in MUFs may be up to 10 degrees above the mid-latitude mid-day median value of 16 MHz. Look for transequatorial openings at that time as well as around December 5 and New Year's Day, when geomagnetic disturbances are expected.

The first, second, and fifth weeks are forecast to have lower solar flux and consequently lower activity. Therefore the lower frequency bands should be excellent then. Disturbances, of course, can provide DX signals from some unusual locations from late in the day to early the next morning — i.e., 10 pm to 4 am local time.

The Geminids meteor shower, which will peak on December 13-14, will provide the richest and most reliable display of the year, with rates of 60 to 70 per hour. Because optical observations may be difficult or impossible during periods of poor weather in December, actual numbers must be determined by radio reception. A smaller version of the shower will be noted on December 22. Lunar perigee will occur on the 2nd and 30th of December and a full moon on December 24th. Winter solstice occurs on December 22nd at 0402 UT.

band-by-band summary

Ten, twelve, and fifteen meters, the day-only DX bands, will be open from morning to early evening almost every day, and to most areas of the world. The openings on the higher of these bands will be shorter, occur closer to

GMT	PST	WESTERN USA N	NE	E	SE	S	SW	W	NW	MST	MID USA N	NE	E	SE	S	SW	W	NW	CST	EST	EASTERN USA N	NE	E	SE	S	SW	W	NW
0000	4:00	40	40	20	12	12	10	10	20	5:00	30	40	20	20	12	10	10	20	6:00	7:00	40	40	20	20	20	15	15	40
0100	5:00	30	40	20	20	12	10	10	20	6:00	40	40	30	20	20	12	12	20	7:00	8:00	40	40	30	20	20	20	20	40
0200	6:00	30	40	20	20	12	12	12	20	7:00	40	40	30	20	20	12	12	30	8:00	9:00	40	80	30	20	20	20	20	40
0300	7:00	40	40	30	20	20	15	15	30	8:00	40	40	30	20	20	15	15	40	9:00	10:00	40	80	30	20	20	20	30*	40
0400	8:00	40	40	30	20	20	20	20	40	9:00	40	40	30	20	20	20	20	40	10:00	11:00	40	80	30	20	20	30	30	40
0500	9:00	40	40	30	30	20	20	20	40	10:00	40	40	30	30	20	20	20	40	11:00	12:00	40	80	30	30	30*	30	30	40
0000	10:00	40	40	30	30	20	20	30*	40	11:00	40	40	30	30	30	20	20	40	12:00	1:00	40	80	30	30	30	30	30	40
0700	11:00	40	40	30	30	20	30	30	40	12:00	40	80	30	30	30	30	30	40	1:00	2:00	40	80	30	30	30	30	30	40
0800	12:00	40	40	30	30	20	30	30	40	1:00	40	80	30	30	30	30	30	40	2:00	3:00	40	80	30	30	30	30	30	40
0900	1:00	40	40	30	30	30	30	30	40	2:00	40	80	30	30	30	30	30	40	3:00	4:00	40	80	30	30	30	30	30	40
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1200	4:00	40	40	30	30	40*	30	30	40	5:00	40	40	15	20	30	30	30	40	6:00	7:00	30	20	12*	15	30	30	30	40
1300	5:00	40	40	20	20	30	30	30	40	6:00	40	30	12	15	30	30	30	40	7:00	8:00	40	20	10	12	20	20	20	40
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1700	9:00	40	30	10	10	15	20	30	40	10:00	40	20	10	10	15	20	20	40	11:00	12:00	40	20	10	10	12	20	30*	40
1800	10:00	40	30	10	10	15	15	20	40	11:00	40	30	10	10	15	15	30	40	12:00	1:00	40	30	10	10	12	15	20	40
1900	11:00	40	40	10	10	12	12	20	40	12:00	40	40	10	10	12	15	20	40	1:00	2:00	40	40*	10	10	12	12	20	40
2000	12:00	40	40	12	10	12	12	15	30	1:00	40	40	12	10	12	12	15	40	2:00	3:00	40	40	12	10	12	12	15	40
2100	1:00	40	40	12	10	12	10	15	20	2:00	40	40	15	10	12	12	15	30	3:00	4:00	40	40	15	10	12	10	12	30
2200	2:00	40	40	15	10	12	10	12	20	3:00	40	40	20	10	12	10	12	20	4:00	5:00	40	40	20	10	12	12	12	20
2300	3:00	40	40	20	10	12	10	12	20	4:00	40	40	20	10	12	10	12	20	5:00	6:00	40	40	20	20	12	12	12	30
	DECEMBER	ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN			ASIA FAR EAST	EUROPE	S. AFRICA	CARIBBEAN S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN

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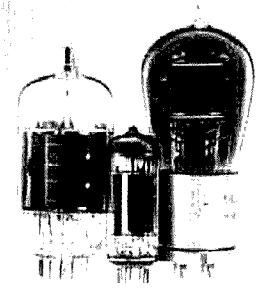
AFFIX LABEL HERE

local noon, and provide paths mainly to the southern hemisphere, with a possibility of transequatorial openings.

Twenty, thirty, and forty meters are both day and night bands. Twenty will be the maximum usable band for DX in the northern directions these days during the daytime. Throughout the night, it will team up with 30 meters. Forty meters will become the main over-the-pole DX daytime band, with some hours covered by 30. This path may be affected by anomalous absorption during a few days of the month.

Eighty and one-sixty meters, the night-only DX bands, will exhibit short-skip propagation during daylight hours, then lengthen for DX at dusk. These bands follow the darkness path, opening to the east just before local sunset, swinging more to the north-south near midnight, and ending up in the Pacific areas during the hour or so before dawn. Eighty is the maximum usable band for some night hours now during the sunspot minimum period; consequently, signal strength and signal quality can be expected to improve. Remember the DX windows of 3790-3800, 1825-1830, and 1850-1855 kHz.

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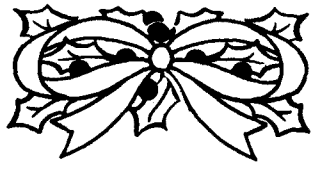
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NEW products

antenna polarity switcher model APS-1

The APS-1 from Mirage is a self-contained control head designed to allow remote polarity switching of circular antennas such as the Mirage/KLM range of crossed Yagis.

The APS-1 may be powered by the power adapter (included) or may alternately be powered from a vehicle or other 13- to 17-volt dc source.

In addition to switchable outputs for two antennas, the APS-1 also contains a 6- to 13-volt regulated dc power supply. This feature is designed for powering items such as preamplifiers and VHF/UHF converters, but may also be used whenever a low-current, stabilized variable voltage source is required.

Total output current is 500 mA with the ac transformer (included), 1 amp with optional high-current transformer or external dc supply.

For information, contact Mirage, P.O. Box 1000, Morgan Hill, California 95037.

Circle #307 on Reader Service Card.

new CTCSS encoder

Kenwood has added a programmable CTCSS encoder to its Smallest HT™. The new BT-Series accessory lets you access *more "private line"* machines with different PL tones.

The introduction of the new BT-Series doesn't mean that the original Smallest HT™ will be discontinued; users who don't need the CTCSS encoder can choose the TH-216A-Series without the front panel programmable tone encoder.

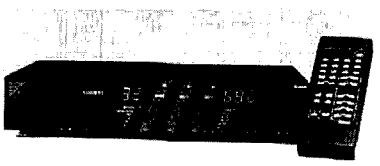
The new BT-Series retains all the same features and uses the same accessories as the original Smallest HT™ and adds a valued feature in the same "pocket-portable package."

For information, contact Trio-Kenwood Communications, P.O. Box 7065, Compton, California 90224.

new Ku-band satellite tv package

Luxor has announced a new Ku-band package for viewers who want to receive both Ku-band and C-band satellite television programming. The Luxor package consists of a Luxor 9995 C/Ku Block Satellite Receiver, a new 1.2- or 1.8-meter fiberglass antenna, choice of mounts, a 9769 Ku-band Low Noise Block Downconverter (LNB), and dual-mode signal feed. The LNB converts the 11.7-12.5 GHz band to 950-1,750 MHz, the input frequency of Luxor block receivers. Its block signal conversion technology permits independent channel selections on TV sets throughout home, building, or neighborhood while sharing one TV antenna.

The 8003 1.2-meter dish and 8004 1.8-meter dish are available with fixed (azimuth elevation)



mounts for commerical use or polar mounts for consumer use.

The Luxor 9769 microwave head consists of the low noise pre-amplifier, mixer, local oscillator, and i-f amplifier, combined in one unit to form the LNB. This LNB (noise factor 2.5 dB maximum), together with the new Luxor 1.2 meter dish, gives an excellent picture on the newer 12 GHz satellites. For reception at footprint edges where signal strength is weak, Luxor recommends its 1.8 meter dish.

For details, contact Luxor North America Corporation, 600 108th Avenue N.E., Suite 539, Bellevue, Washington 98004.

Circle #306 on Reader Service Card.

antenna current probe

Antenna experimenters will find the new MFJ-206 antenna current probe to be a handy addition to their hamshack. Designed to detect magnetic fields around rf conductors such as antennas, transmission lines, and ground leads, it can also be also used to detect electromagnetic radiation from wiring, enclosure shields, and other objects. It can also be used as a field-strength meter with the included telescoping antenna. The MFJ-206 is priced at $79.95.

The relative strength of a magnetic field is determined by the magnitude of the rf currents flowing through a conductor. To detect current distribution, you simply move the slotted end of the probe along the conductor at right angles. As current increases, you'll see a higher reading on the probe's meter.

For further information, contact MFJ Enterprises, Box 494, Mississippi State, Mississippi 39762.

Circle #305 on Reader Service Card.

SSB voice-activated squelch

Naval Electronics announces a new commercial-grade voice-operated squelch (VOS) for upgrading solid-state SSB receivers and transceivers.

Easily installed in most modern receivers and transceivers, the 11-pin module provides commercial communications-grade squelch operation immune to noise, static crashes, heterodynes, and steady tones. The circuit incorporates a syllabic rate detector which provides for superior squelch operation, looks for signal changes similar to changes normally occurring in human speech, and uses the VOS to switch the speaker audio on. (An on/off switch is provided.)

The VOS is most helpful where a single channel must be monitored continuously, reducing operator fatigue.

For further information, contact Naval Electronics Inc., 5417 Jetview Circle, Tampa, Florida 33634.

Circle #310 on Reader Service Card.

digital oscilloscope peripheral for pc's

Rapid Systems has announced the release of the new "4x4," a digital oscilloscope peripheral for IBM PC, XT, and ATs and compatibles. This turn-key, high-performance, 128K data buffer digital scope is the first serious instrument peripheral for PC's to offer simultaneous acquisition of four channels. It also offers the largest number of data buffers per channel in the market. Turn-key digital scope software is provided with unit. No programming is required.

For information contact Rapid Systems, Inc. 755 North Northlake Way, Seattle, Washington 98103.

Circle #304 on Reader Service Card.

free update disks

Free update discs are available to anyone who's bought *RF Notes 1* (through Version 2.4), *RF Notes 2* (through Version 2.4), or *RF Notes 3*, Volume 1 (through Version 1.5).

For a free copy, simply return your original program disk to Etron RF Enterprises, P.O. Box 4042, Diamond Bar, California 91765.

Circle #303 on Reader Service Card.

new DX Edge®

There's a new DX Edge from Xantek – the Super DX Edge™, the successor to the "Computerized DX Edge." Available for the Commodore 64 and Commodore 128 computers and their associated disk drives, the Super DX Edge adds two significant features not found in earlier versions: it calculates Maximum Usable Frequency (MUF) between any two locations and calculates Great Circle Bearings (antenna direction) and distance between any two locations.

The program calculates the MUF between any two points for any sunspot number or solar flux value, and also the great circle bearing between the points. The MUF is the frequency to use for most effective communications between any two points. Results are given for every hour of the day.

The super DX Edge displays and automatically updates the position of the Gray Line to give a continuously accurate picture of the world's daylight and darkness areas. Sunrise and sunset times for any location and any time of year may

also be calculated and displayed.

The DX Edge is used by DXers around the world to help determine the optimum times for long-distance contacts. It is particularly helpful in determining the best times and bands for long path and Gray Line contacts. With the addition of the MUF and Great Circle Calculators, the Super DX Edge contains all the propagation aids that DXers need to plan their on-the-air activity.

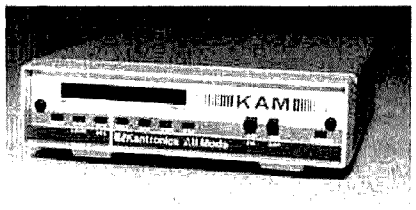
Available from Ham Radio's Bookstore or directly from Xantek, Super DX Edge is priced at $34.95. (Add $3.50 for shipping and handling if ordering from HR's Bookstore.) Owners of the Computerized DX EDGE can trade up to the Super DX EDGE for $24.95 plus their disk.

For information, contact Xantek, Inc., P.O. Box 834, Madison Square Station, New York, New York 10159.

Circle #302 on Reader Service Card.

all-mode communicator

The KAM™ is a true all-mode digital "smart" communicator available from Kantronics. A state-of-the-art radio modem that operates CW, RTTY, ASCII, AMTOR, or PACKET, KAM



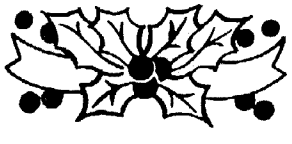
features over 100 software commands, bar graph tuning and user-programmable MARK and SPACE tones for RTTY and hf packet, RS-232/TTL terminal interfacing, and limiter/limiterless operation on hf for weak-signal operation. In addition, the CW demodulator is programmable for both center frequency and bandwidth. In effect, you get a flexible, all-mode unit, with any shift — including 170 Hz for RTTY — without compromise.

Like a packet unit, KAM operates with a command mode and operating modes. You can choose RTTY or packet by command. In addition, you can set mark and space tones for RTTY or packet while in command mode for hf operation. Standard packet tones are used in the VHF mode. You'll need a terminal program to access KAM with you computer; any standard communications program will do.

KAM comes with a power adapter, an operator's manual, cables, and connectors; you provide the transceiver mic connectors and your computer's RS-232 connector.

For details, contact Kantronics, 1202 E. 23 Street, Lawrence, Kansas 66046.

Circle #301 on Reader Service Card.



1987 CALLBOOKS

147



The "Flying Horse" sets the standards

Continuing a 66 year tradition, there are three new Callbooks for 1987.

The North American Callbook lists the calls, names, and address information for licensed amateurs in all countries from Canada to Panama including Greenland, Bermuda, and the Caribbean Islands plus Hawaii and the U.S. possessions.

The International Callbook lists the amateurs in countries outside North America. Coverage includes South America, Europe, Africa, Asia, and the Pacific area.

The 1987 Callbook Supplement is a new idea in Callbook updates; it lists the activity in both the North American and International Callbooks. Published June 1, 1987, this Supplement will include all the new licenses, address changes, and call sign changes for the preceding 6 months.

Publication date for the 1987 Callbooks is December 1, 1986. See your dealer or order now directly from the publisher.

☐ North American Callbook
Incl. shipping within USA $28.00
Incl. shipping to foreign countries 30.00

☐ International Callbook
Incl. shipping within USA $28.00
Incl. shipping to foreign countries 30.00

☐ Callbook Supplement, published June 1st
Incl. shipping within USA $13.00
Incl. shipping to foreign countries 14.00

SPECIAL OFFER

☐ Both N.A. & International Callbooks
Incl. shipping within USA $53.00
Incl. shipping to foreign countries 58.00

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All payments must be in U.S. funds.

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Dept. F
925 Sherwood Dr., Box 247
Lake Bluff, IL 60044, USA

Tel: (312) 234-6600 VISA

126

Crystal Filters

YEAR-END CLEARANCE

All starred (★) items 20% off; all others 10%. Prices are *each* except as noted. All filters 8-pole. Sale ends December 31, 1986.

FILTERS FOR KENWOOD - Reg. $60 except as noted.
8.83MHz IF for models: TS120 through TS940
Bandwidths: 250, 400, 1800, ★ 2100, 6000Hz
TS440 ★ Pair (400 Hz CW, 2.1KHz SSB) . . . . Reg. $120
TS430 ★ Triple (Both above plus AM) . . . . . Reg. $180

455 KHz IF for R820, TS830/930/940 . . . . . Reg. $110
Bandwidths available: CW 400HZ; SSB ★ 2.1KHz

Matched Filter Pairs for Above . . . . . Reg. $170 pr.
(8.83mhZ and 455KHz) SSB: ★ 2100Hz, CW: 400Hz

3.395MHz IF for TS520, TS511, R599.
Bandwidths Available: 250, 400, ★ 1600, 2400Hz

Filter Cascade Kits with Filter and Amplifier
For ★ TS430 $85, ★ TS520 $80, ★ TS820 — Reg. $80

FILTERS FOR YAESU . . . Reg $60 except as noted.
3.18MHz IF for FT-101 Series except Z/D.
BWs: 250, 500Hz, 1.8, ★ 2.1, ★ 2.4, ★ 6.0KHz

8.2 MHz IF for FT-102, FT-757/767.
Bandwidths Available: ★ 250, 500, 2100Hz
454KHz IF for FT-102 (★ 250, 500Hz) . . . . . Reg. $75
455KHz IF for FT-102 (★ 2.1KHz) . . . . . Reg. $110

8.9MHz for FT-101ZD/107/707/901-2, FT-980.
BWs: 250, 500Hz, 1.8, 2.1, ★ 2.4, 6.0KHz
10.76MHz IF for all but 980: BWs: 1.8, ★ 2.1KHZ
455 IF for FT980 only: BW ★ 2.1KHz . . . . . Reg. $110
455.8 IF for FT-980, FT726: BW ★ 500Hz . . . Reg. $75

Filter Cascade Kits with Filter and Amplifier
For ★ all 8.9 above except FT-980 . . . . . Reg. $80

9.0MHz IF for Tempo I (or FT-200), FT-301, FT-7/B
BWs: ★ 250, 500Hz, 1.8, ★ 2.1, 2.4, 6.0KHz
NOTE: Above are our "homebrewers' favorites"!

FILTERS FOR ICOM (exact replacements)
455 IF for IC730/740/745/751, R70/71, etc.
Bandwidths: FL44A (SSB ★ 2.4KHz) . . . . . Reg. $110
FL52A (500Hz); FL532 (250Hz) . . . . . Reg. $85 ea.

FILTERS FOR HEATH - ALL MODELS . . . . . Reg. $65
Bandwidths Available: 250, 400Hz ★ 1.8, 2.1KHz
For SB-104 Only: ★ 400HZ (3395.7 IF)

FILTERS FOR DRAKE R-4C . . . Reg. $65 exc as noted
GUF1 (8KHz) Replaces original 1st IF 4-pole unit
GUF2 (800Hz) Switches out GUF1 in CW . . . Reg. $100
2nd IF 125 ($75), 250, 400Hz, 1.8, ★ 2.1, 6KHz

FILTERS FOR DRAKE TR7/R7, etc. . . . . Reg. $65
BWs Available: 250, 400Hz; ★ 1.8, ★ 2.1KHz

FILTERS FOR COLLINS 75S-3B/C ★ 250Hz Reg. $125

LIMITED QUANTITIES - ORDER NOW!
SPECIFY: Make and Model Number of your Rig.
Frequency and Bandwidth of filter(s)
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SHIPPING: $5 US and Canada, $12 elsewhere.

GET THE BEST 8-POLE FILTERS — FOR *LESS*

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SOFTWARE FOR THE C-64 *

RF KIT 1.0 disk includes over a dozen RF and DC formulas, programs for Beam Headings, Tower Stress Calc and Noon Meridian to find True North. Also, instructions to add your own favorite formulas to the disk. Write for free data. Price $14.95 + 1.00 postage.

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127

RF KIT CO., P.O. Box 27127, Seattle WA 98125 USA
Tel (206) 246-9522 *Trademark of Commodore

NEMAL ELECTRONICS

HARDLINE — 50 OHM

Nemal No.	Description	Per Ft.
FXA12	1/2" Aluminum Black Jacket	89
FLC12	1/2" Corr Copper (EQ Heliax® LDF4) Blk Jkt	1 59
FLC78	7/8" Corr Copper	3 92
NM12AL	N Conn 1/2" Alum (Male or Female)	22 00
NM12CC	N Conn 1/2" Copper (Male or Female)	22 00
NM78CC	N Conn 7/8" Copper (Male or Female)	54 00

COAXIAL CABLES

Nemal No.	Description	100 Ft.	Per Ft.
1100	RG 8 95% Shielded Mil Spec	28 00	32
1102	RG8 95% Shielded Foam	30 00	32
1110	RG8X 95% Shield (mini 8)	15 00	17
1130	RG213/U Mil Spec 96% Shield	34 00	36
1140	RG214/U Mil Spec Dbl Silver	155 00	1 65
1180	Belden 9913 Low Loss	46 00	50
1705	RG142B/U Teflon/Silver	140 00	1 50
1310	RG217/U 5/8" 50 ohm Dbl Shield	80 00	85
1470	RG223/U Mil Spec Dbl Silver	80 00	85
1450	RG174 95% Shielded Mil Spec	12 00	14

ROTOR CABLE — 8 COND.

Nemal No.	Description	100 Ft.	Per Ft.
8C1822	2 18 Ga 6 22 Ga	19 00	21
8C1620	2 16 Ga 6 20 Ga Heavy Duty	34 00	36

* Shipping $3 00 – 100 Ft / Conn $3 00 / C O D $2 00

CONNECTORS — MADE IN U.S.A.

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NE720	Type N for Belden 9913		4 25
NE723	N Female Belden 9913		4 75
PL258AM	Amphenol Barrel		1 45
PL259	Standard Plug for RG8 213	10/5 90 or	65
PL259AM	Amphenol PL259	10/7 90 or	89
PL259IS	PL259 Teflon/Silver		1 59
UG21D	Type N for RG8 213 214		3 00
UG83B	N Female to PL259		6 50
UG88C	BNC RG58		1 25
UG146	SO239 to Male N		6 50
UG175/6	Adapter for RG58/59 (specify)	10/2 00 or	22
UG255	SO239 to BNC Amphenol		3 75
KA51-18	TNC RG58		4 35
AM9501 1	SMA RG142B		8 95
SO239AM	Amphenol SO239		89

GROUND STRAP — BRAID

Nemal No.	Description	Per Ft.
GS38	3/8" Tinned Copper	30
GS12	1/2" Tinned Copper	40
GS316	3/16" Tinned Copper	15
GS316S	3/16" Silver Plated	35

GROUND WIRE — STRANDED

Nemal No.	Description	Per Ft.
HW06	6 Ga insulated stranded	35

Call or write for complete price list. Nemal's 32-page Cable & Connector Selection Guide is available at no charge with orders of $50.00 or more, or at a cost of $4.00 individually.

NEMAL ELECTRONICS, INC. (305) 893—3924 • Telex 6975377
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Outstanding performance of W9INN antennas is well known! Now on-for multi-band DIG-SIGNAL reports! Automatic bandswitching • Very low SWR • Coax feed • 3kw power • Compact • FULLY ASSEMBLED to your specified center frequency each band • EASY TO INSTALL • Very low profile • Complete instructions • Your personal check accepted

4 BAND SLOPER	160, 80, 40, 30, or 20M	60 ft long	$ 48 ppd
3 " "	160, 80, 40M	60 ft "	$ 43 "
2 " "	80, 40M	40 ft "	$ 35 "
3 " NO-TRAP DIPOLE	160, 80, 40M	133 ft long	$ 71 "
2 " " "	80, 40M	65 ft "	$ 55 "
9 BAND SPACE-SAVER DIPOLE*	160 thru 10M	46ft long	$ 85 ppd

* Requires wide-range tuner (80, 40, 20, 15M without tuner)

SEND SASE for complete details of these and other unique antennas

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1986-87 CALL DIRECTORY
(on microfiche)

Call Directory	$8
Name Index	$8
Geographic Index	$8

All three — $20
Shipping per order $3 ↙ 122

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"Impressively No. 1 Receiver... Way ahead... Best buy." R. Lichte
JRC NRD 525... $1179 (+$9 s&h)



Successor to the famous professional grade NRD 515, this new model incorporates much of its predecessor and adds many new features of its own to offer truly quality performance.

- Microprocessor controlled tuning, mode select, triple scanning, memory and clock functions
- Tunes 150 kHz to 29.999 MHz
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- Keyboard frequency entry
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- Squelch
- Passband tuning
- Incremental tuning
- Attenuator
- Tone control
- Programmable clock/timer
- Sensitivity: (10 dB S/N) .09-1.6 MHz, RTTY, FAX, CW, SSB: 5 µV; AM 15 µV. 1.6-34 MHz, RTTY, FAX, CW, SSB: 0.5 µV; AM 2 µV.
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For Technical Review of the 525 by Rainer Lichte Send $3.00

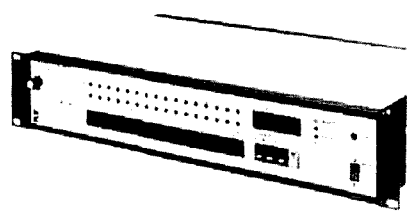
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↙ 123

NEW products

"talking" DTMF remote control

The Model 5001 DTMF Remote Control from Monroe Electronics answers DTMF-coded inquiries and generates alarm reports in English using synthesized speech. The device monitors 16 on-off type logic inputs and controls 16 outputs through built-in Form C relays. Communication with the unit is by means of dual tone multiple-frequency (DTMF) signals from any touch-tone telephone or over any other voiceband link equipped with a touch-tone keypad.



The Model 5001 controls its output relays as instructed, and reports the state of its logic inputs when queried by the user. It will generate an alarm report spontaneously when designated inputs change state, autodialing one of two preprogrammed telephone numbers or transmitting via two- or four-wire audio ports. User-selectable protocols for status reporting include synthesized speech, DTMF tone strings, DTMF tones, and single tone signals.

The 5001 accepts instructions via the public, switched telephone network. It answers after a user-designated number of rings, and hangs up automatically after a selectable period of inactivity. Relays may be programmed for a variety of configurations. Applications include repeater and satellite uplink/downlink control.

For information, contact Monroe Electronics, Inc., 100 Housel Avenue, Dept. 555, Lyndonville, New York 14098.

Circle #308 on Reader Service Card.

prime factor FFT

Alligator Transforms has announced the release of Prime Factor FFT, a Fast Fourier Transform Subroutine Library usable with the IBM PC, XT, AT, and Compatibles equipped with an 8087 *math coprocessor. The library includes the* forward and inverse FFT for single and double precision floating point complex number sets.

Prime Factor FFT may be called from any high-level language including Turbo Pascal, Fortran, BASIC, C, or Assembly Language. Interface ex-

amples are provided in all languages with extensive documentation on how to use Fast Fourier Transforms in your application.

With Prime Factor FFT, the user is not limited to radix two data set sizes. You may perform frequency analysis on data set sizes as small as two data points to as large as 65,520 data points. Choose from over 300 data set sizes.

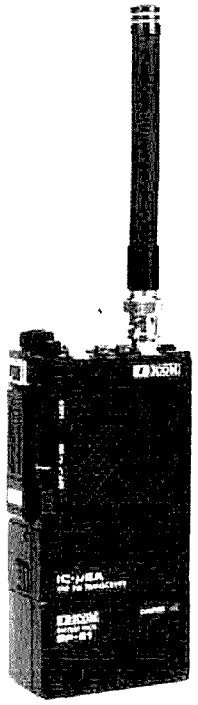
Prime Factor FFT fully utilizes the 8087 math coprocessor to complete calculations quickly. A 1008-point double precision floating point complex forward FFT will complete in 2.428 seconds. Larger data sets execute quickly as well with a similar 32,760 data point set FFT completion in 133.863 seconds. Priced at $159.00 each, Prime Factor FFT is extremely accurate, with typical error levels as low as 3.3307e-16 in double precision.

For information, contact Alligator Transforms Scientific Software, P.O. Box 11386, Costa Mesa, California 92627

Circle #309 on Reader Service Card.

ICOM IC-μ2AT 2-meter handheld

ICOM has announced the release of its pocket-sized 2-meter handheld, the IC-μ2AT. The "Micro" receives from 139 to 174 MHz and transmits from 140 to 150 MHz, has ten memories which will store offset and tone, an LCD readout on the top panel, scanning, 1-W power output (with 1.5W optional), and 32 built-in subaudi-

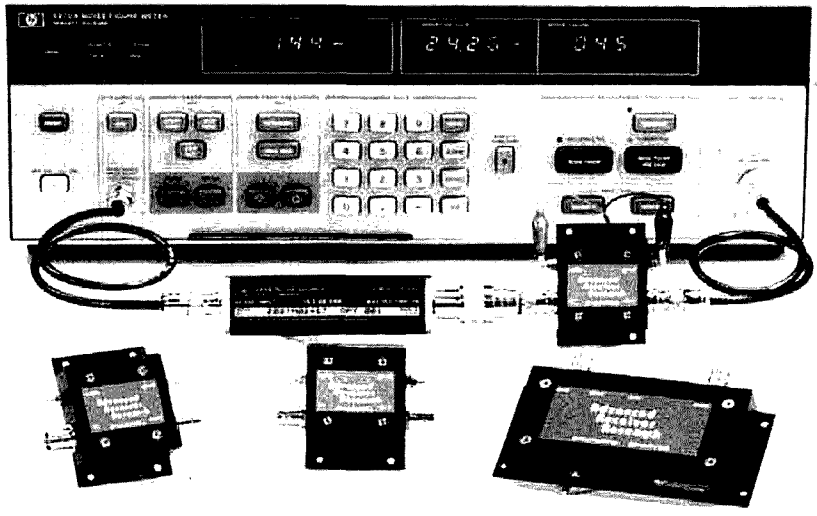


ble tones. It measures only 2.3 × 5.6 × 1.1 inches and weighs only half a pound.

For details, contact ICOM America Inc., 2380 116th Avenue N.E., Bellevue, Washington 98009-9029.

Circle #311 on Reader Service Card.

High Performance vhf/uhf preamps



Receive Only	Freq. Range (MHz)	N.F. (dB)	Gain (dB)	1 dB Comp. (dBm)	Device Type	Price
P28VD	28-30	< 1.1	15	0	DGFET	$29.95
P50VD	50-54	< 1.3	15	0	DGFET	$29.95
P50VDG	50-54	< 0.5	24	+ 12	GaAsFET	$79.95
P144VD	144-148	< 1.5	15	0	DGFET	$29.95
P144VDA	144-148	< 1.0	15	0	DGFET	$37.95
P144VDG	144-148	< 0.5	24	+ 12	GaAsFET	$79.95
P220VD	220-225	< 1.8	15	0	DGFET	$29.95
P220VDA	220-225	< 1.2	15	0	DGFET	$37.95
P220VDG	220-225	< 0.5	20	+ 12	GaAsFET	$79.95
P432VD	420-450	< 1.8	15	- 20	Bipolar	$32.95
P432VDA	420-450	< 1.1	17	- 20	Bipolar	$49.95
P432VDG	420-450	< 0.5	16	+ 12	GaAsFET	$79.95
Inline (rt switched)						
SP28VD	28-30	< 1.2	15	0	DGFET	$59.95
SP50VD	50-54	< 1.4	15	0	DGFET	$59.95
SP50VDG	50-54	< 0.55	24	+ 12	GaAsFET	$109.95
SP144VD	144-148	< 1.6	15	0	DGFET	$59.95
SP144VDA	144-148	< 1.1	15	0	DGFET	$67.95
SP144VDG	144-148	< 0.55	24	+ 12	GaAsFET	$109.95
SP220VD	220-225	< 1.9	15	0	DGFET	$59.95
SP220VDA	220-225	< 1.3	15	0	DGFET	$67.95
SP220VDG	220-225	< 0.55	20	+ 12	GaAsFET	$109.95
SP432VD	420-450	< 1.9	15	- 20	Bipolar	$62.95
SP432VDA	420-450	< 1.2	17	- 20	Bipolar	$79.95
SP432VDG	420-450	< 0.55	16	+ 12	GaAsFET	$109.95

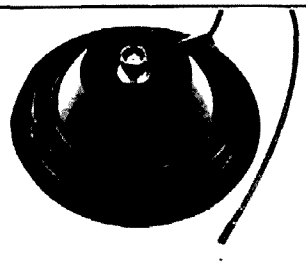
Every preamplifier is precision aligned on ARR's Hewlett Packard HP8970A/HP346A state-of-the-art noise figure meter. RX only preamplifiers are for receive applications only. Inline preamplifiers are rf switched (for use with transceivers) and handle 25 watts transmitter power. Mount inline preamplifiers between transceiver and power amplifier for high power applications. Other amateur, commercial and special preamplifiers available in the 1-1000 MHz range. Please include $2 shipping in U.S. and Canada. Connecticut residents add 7-½% sales tax. C.O.D. orders add $2. Air mail to foreign countries add 10%. Order your ARR Rx only or inline preamplifier today and start hearing like never before!

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MAGNET MOUNTS

For HT owners operating inside a vehicle and wanting increased T/R range, RF PRODUCTS has the low cost solution.

Remove your BNC antenna from the HT and mount on the RF PRODUCTS BNC magnet mount. Install the magnet mount on the roof top and connect the BNC co-ax connector.

The magnet mount (part no. 199-445) has 10 feet of small (5/32") co-ax with BNC conector attached. **PRICE** $15.95 M.O. or cashiers ck., via UPS gnd. Fla. residents add 5% tax, for air UPS add $3.25

The RF PRODUCTS Magnet Mounts are one of the few mounts available that can be repaired should the co-ax cable be damaged. The large surface area capacitance disc provides proper ground plane coupling for 1/4 and 5/8 wavelength VHF and UHF antennas.
MODELS AVAILABLE WITH THE FOLLOWING CONNECTORS & CO-AX TYPES.
ANTENNA CONNECTORS: BNC, TNC, 1 1/8"(MOT.), 5/16-24 STUD, 3/8-24 SOCKET.
CO-AX CABLE: RG-122/U, RG-58A/U, mini 8X.
TRANSCEIVER CONNECTORS: BNC, TNC, PL-259, type N.

125

RF PRODUCTS
P.O. Box 33, Rockledge, FL 32955, U.S.A. (305) 631-0775

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PACKET PROGRAM FOR YOUR PC. A telecommunications program especially written to interface your packet controller with your PC or compatible. Features Split-Screens, ASCII, Binary, X-Modem file transfers. DOS shell. Macro Keys. 5 types *time/date/oper. ID. Lots of goodies. Software support. Write* for complete information. VISA/MC. Kalt and Associates, 2440 E. Tudor Rd, Suite #138, Anchorage, AK 99507. (907) 248-0133.

REMEMBER TROLLEY CARS? Trolley Treasures: "The Wartime Years in New Jersey", a 4-volume photodocumentary history, includes 1600 unpublished, original photographs plus extensive historical notes. Volume I, "The Compromise Roof Cars of Public Service Coordinated Transport", ready now. For details, send SASE to Trolley Themes, A.W. Mankoff, 2237-3 Woodside Lane, Sacramento, CA 95825. Or order directly from publisher ($12.95 plus $1.50 S&H): Railhead Publications, PO Box 526HR, Canton, Ohio 44701.

NATIONAL COLLECTORS: National HFS rcvr, all original tubes, coils and power supply. Very good condition. Make offer. Ray Nether, W3KFB, 184 McClure road, RD 2 Cheswick, PA 15024.

1296 MHz POWER AMPLIFIERS. 6 W Kit—$80.65- $3.50 s/h. 20 W Kit—$112.15—$3.50 s/h. For info SASE to: A & A Engineering, 2521 W. La Palma, -K, Anaheim, CA 92801. (714) 952-2114.

UHF PARTS. GaAs Fets, MMICs, trimmers, chip caps, teflon pcb, and other builder parts. SASE brings list. Microwave Components, 11216 Cape Cod, Taylor, MI 48180.

SHOW IT IN STYLE. Full color QSL's by Smith Printing. From your prints/slides. Sample packet. SASE. 20420 Calhaven Drive, Saugus, CA 91350. (805) 251-7211.

LOOKING FOR the following equipment not modified: ARC 4—TBY—APS13—BC645 or other items used in aviation in the years 1939-45. Giuliano Vigarani, I4VGG.

WORLDWIDE AIR TRAFFIC CONTROL Frequency Directory for 2-22 MHz shortwave aeronautical bands. Hear commercial airliners on oceanic and many foreign domestic routes. Cross-referenced lists include over 200 cities plus weather and airline channels. Send $4.75 (Mass residents add 5%) + $1.25 P&H ($2.25 outside N. America) to Cambridge Airadio, Dept HRA, Suite 486, 89 Mass. Avenue, Boston, MA 02115.

BUG COLLECTORS: Original Bunnell Gold Bug, S/N 715. Also have Bunnell-Martin Flash Key, Type No. 5-46. Make offer. Ray Nether, W3KFB, 184 McClure Road, RD 2, Cheswick, PA 15024.

TRS80 4P/KANTRONICS UTU RTTY. Split-screen, 10 user keys, file transfer. Runs in Mod 4 (80 char) mode, $30 to COMMPRO *RTTY, c/o KB6IC, 3711 Gayle Avenue, Omaha, NE* 68123.

SPECIAL ANTENNAS by K3IPW featuring the RIW-19 432 MHz beam, 14.9 dBd for $69.95 and the K1FO 22 el. 14 ft. 432 MHz beam, 15.8 dBd for $76.64, SASE for information. Thomas H. Rutland, 1703 Warren St, New Cumberland, PA 17070.

GENERAL RADIO TEST EQUIPMENT 1211B unit oscillator 0.5 to 50 MHz $35. Two 1209B unit oscillators 250 to 920 MHz $30 each. 1201B regulated unit power supply $20. 1203B unit power supply $15. All work. All for $100. Shipping extra. Victor Smith, 147-29 Hoover Avenue, Briarwood, NY 11435.

$$$$$SUPER SAVINGS on electronic parts, components, supplies and computer accessories. Free 40-page catalog for SASE. Get on our mailing list. BCD ELECTRO, PO Box 830119, Richardson, TX 75083 or call (214) 690-1102.

CABLE TV CONVERTERS & EQUIPMENT: Plans and parts. Build or buy. SASE for information. C & D Electronics, PO Box 1402, Dept. HR, Hope, AR 71801.

YAESU OWNERS: Hundreds of modifications and improvements for your rig. Select the best from fourteen years of genuine top-rated Fox-Tango Newsletters by using our new 32-page *Cumulative Index.* Only $5 postpaid (cash or check) with $4 Rebate Certificate creditable toward Newsletter purchases. Includes famous Fox-Tango Filter and Accessories Lists. Milt Lowens, N4ML (Editor), Box 15944, W. Palm Beach, FL 33416. Telephone (305) 683-9587.

R-390A Receiver: $195 checked; $115 reparable. Parts, tubes, sections. Info SASE. CPRC-26 six meter transceiver (see HR, March 1985) $17.50 apiece, $32.50 pair (add $4.50/unit shipping). Baytronics, Box 591, Sandusky, OH 44870. 419-627-0460 evenings.

MARCO: Medical Amateur Radio Council, Ltd. operates daily and Sunday nets. Medically oriented Amateurs (physicians, dentists, veterinarians, nurses, physiotherapists, lab technicians, etc.) invited to join. Presently over 550 members. For information write MARCO, Box 73's, Acme, PA 15610.

IBM-PC RTTY/CW. New CompRtty II is the complete RTTY/CW program for IBM-PC's and compatibles. Now with *larger buffers,* better support for packet units, pictures, much more. Virtually any speed ASCII, BAUDOT, CW. Text entry via built-in screen editor! Adjustable split screen display. Instant mode/speed change. Hardcopy, diskcopy, break-in buffer, select calling, text file transfer, *customizable full screen logging,* 24 programmable 1000 character messages. Ideal for MARS and traffic handling. Requires 256k PC or AT compatible, serial port, RS-232C TU $65. Send call letters (including MARS) with order. David A. Rice, KC2HO, 25 Village View Bluff, Ballston Lake, NY 12019.

ELECTRONIC ENCLOSURES: Attractive, reasonably priced. Aluminum top and sides. Walnut stained solid wood ends. Variety of sizes and custom sizes available. Sample $5.00 Information free. Energy Engineering, Custom Division, Rt. 4, Box 330, Fayetteville, AR 72701.

NATIONAL RADIO EQUIPMENT manual list or NCL-2000 parts kits. SASE. Maximilian Fuchs, 11 Plymouth Lane, Swampscott, MA 01907.

RTTY JOURNAL—Now in our 34th year. Join the circle of RTTY friends from all over the world. Year's subscription to RTTY JOURNAL, $10.00, foreign $15.00. Send to: RTTY JOURNAL, 9085 La Casita Ave., Fountain Valley, CA 92708.

IMRA International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, New York 10538.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. Daily Electronics, PO Box 5029, Compton, CA 90224. (213) 774-1255.

CUSTOM MADE EMBROIDERED PATCHES. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept 301, 4202 N. Drake, Chicago, IL 60618.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

CABLE TV CONVERTERS/DESCRAMBLERS. Guaranteed lowest prices in US. Jerrold, Hamlin, Zenith—Many others. Lowest dealer prices! Orders shipped within 24 hours! Mastercard, VISA, COD accepted. Free Catalog—call (800) 345-8927 only Pacific Cable Co., Inc., 7325-1/2 Reseda Blvd. #1019, Reseda, CA 91335. (818) 716-5914.

WANTED: Heath ET-100. Call 913-898-4695. Mike W0XM.

MORSE CODE Practice program for IBM-PC and Compatibles. Send $19.95 check to SP Microcomputing Co., 1008 Swallow Drive, Cherry Hill, NJ 08003. Developed by KD2SM.

CHASSIS and cabinet kits. SASE K3IWK, 5120 Harmony Grove Road, Dover, PA 17315.

COMING EVENTS

Activities — "Places to go . . ."

CALIFORNIA: FCC exams, Novice-Extra. Sunnyvale VEC ARC. (408) 255-9000 24 hour. 73, Gordon, W6NLG, VEC

MASSACHUSETTS: The MIT UHF Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday, December 17, 7 PM, MIT Room 1-134, 77 Mass Avenue, Cambridge MA. Reservations requested 2 days in advance. Contact Ron Hoffmann (617) 253-0160/646-1641 or Craige Rodgers (617) 494-1986. Exam fee $4.25. Bring copy of current license (if any), two forms of picture ID and completed form 610 available from FCC in Boston (223-6609)

PENNSYLVANIA: Amateur Radio Brunch. The Chaverim of Delaware Valley will hold their 7th annual brunch, Sunday, January 18, Parktowne Place Apartment Complex, 2200 Benjamin Franklin Parkway, Philadelphia at 10 AM. Everyone invited. Contact Bill Soble, W3QXT (215) 676-6769 or write to 9357 Hoff Street, Philadelphia, PA 19115. Reservations are necessary.

WISCONSIN: The 15th annual Midwinter Swapfest, sponsored by the West Allis Radio Amateur Club, Saturday, January 10, Waukesha County Expo Center Forum, 8 AM to 3 PM. Admission $2.00 advance, $3.00 at the door. 4' tables $3.00 advance, $4.00 at the door. Elec. outlet $5.00. Advance deadline January 2, 1987. Dealers welcome. Amateur exams given, write for details. For tickets or information write WARAC Swapfest, PO Box 1072, Milwaukee, WI 53201. Please SASE.

OPERATING EVENTS

"Things to do . . ."

December 6: The University of Idaho ARC, W7UQ, will hold its 2nd annual Alumni Reunion on the Air, 2000Z, December 6 to 0400Z December 7. All Amateurs, especially U of I alumni, are invited to participate. Listen for "CQ Reunion". QSL available by sending SASE via callbook address. For more information contact W7UQ.

December 6-7: 22nd Annual Pioneer Radio QSO Party sponsored by the John D. Burlie Chapter No. 89. For further information Ted Phelps, W8TP.

Feb 1-2: YL-ISSB QSO Parties. CW 0001Z to 2359Z. Phone 0001Z March 21 to 2359Z March 22. Open to all. U.S. General Class portions of bands. Exchange call, report, QTH, name, ISSB number and partner or teammate. Certificates to category, country/state winners. For information: Bill Early, WA9AEA, PO Box 401, McHenry, IL 60050-0401, USA.

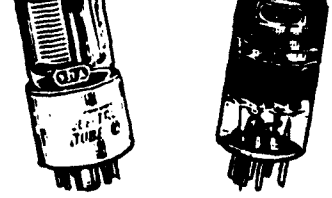
SOUTH BEND, INDIANA Hamfest Swap & Shop, January 4, 1987, first Sunday after New Year's Day, Century Center downtown on US 33 Oneway North between St. Joseph Bank Bldg and river. Four lane highways to door from all directions. Tables: $5/5 ft. Round: $10/8x2.5. Rectangular: $2/ft. wall locations. Talk in freq: 52.52, 99-39, 93-33, 69-09, 145.29. K9IXU (219) 233-5307.

Oecember 12-14: Christmas, Florida. The Coronado Wireless Assn. will operate K4HML from 1400Z to 220Z, lower 10 kcs of General bands, phone and CW. Send QSL and No. 10 SASE to K4HML, PO Box 1, Edgewater, FL 32032.

December 20-21: Christmas City Station Members of the Delaware-Lehigh ARC will run this special event station, W3OK. 1600 UTC to 2200 UTC. Contact the station and send QSL and No. 10 SASE to W3OK, Delaware-Lehigh ARC, Greystone Building, Nazareth, PA 18064.

December 13-14: Bethlehem, IN. Clark County ARC will operate W9WWI/9 from 1700Z to 0300Z December 13 and 1300Z to 2000Z December 14. Certificate for large SASE via CCARC, Box 532, Jeffersonville, IN 47131.

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